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Measurement Techniques

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LINEAR MEASUREMENTS

METHOD FOR COMPLEX SINGLE-PROFILE TESTING OF GEAR WHEELS AT OPERATING LOADS AND SPEEDS

M. A. Skuridin and L. A. Kudryavtsev

Translated from Izmeritel'naya Tekhnika, No. 7,
pp. 1-3, July, 1961

Gear transmission is a system in which mechanical vibrations can easily arise, due to the presence of moving bodies and elastic forces. These vibrations are usually caused by errors in the manufacture of the wheels, distortions of the teeth profiles under stresses, and suddenly changing elasticity of the teeth due to excessively close coupling. The mutual relationship of these causes of vibrations and their effect on the operation of the gear transmission vary considerably with load and speed.

The existing static methods of checking gears only reveal manufacturing errors and completely miss the errors due to deformations and the varying elasticity of the teeth.

At the Moscow Machine-Tool and Instrument Institute, according to the idea and under the guidance of D. Sc. M. A. Skuridin, a test rack* was constructed which provides on an oscillogram the curve for a complex single-profile test of gear wheels under operating loads and speeds. The operational quality of the wheel is checked under conditions which closely approach those encountered in practice, thus eliminating in their evaluation many errors which may arise if only static testing methods are used.

Excessive displacements of the driven gear link due to the above-mentioned causes can be measured with an error of ± 3 angular seconds on this test rack. These excessive displacements are measured on the basis of the torsiograph principle, i. e., with respect to a flywheel whose rotation can be assumed to be uniform.

The loading of a gear transmission is easily varied by means of an electromagnetic brake, and its speed by means of a Ward-Leonard control system.

Figure 1 shows the schematic of the test rack. A common base carries the driving stock 3, the driven stock 13, the electromagnetic powder brake 17, and slip-ring devices 1 and 18. The stock mandrels carry the tested gear wheels 7 and 8.

From driving pulley 2 the torque is transmitted through the driving mandrel to gear 7, the driven gear 8, the journal of the driven shaft and through the tapered pivot to torsion shaft 9, which is placed inside the driven mandrel. The other end of torsion shaft 9 transmits the torque to the shaft of brake 17.

Flywheel 12, with respect to which the excessive angular displacements of the driven gear 8 are measured, is mounted on the driven shaft on precision ball bearings 11, and is connected to the shaft by spring 10. Strain gauges are glued to the spring. The elastic inertial parameters of the system formed by the flywheel and the spring are chosen in such a manner that the natural oscillation frequency of the system is considerably smaller than that of the tested gear transmission oscillations.

The excessive displacements of the driving shaft can also be recorded by means of flywheel 5 with its spring and strain gauge.

In order to provide a linear characteristic for the transducer, spring 10 has a bend in it and carries its strain gauges in a section set at right-angles to the radial portion of the spring, thus making negligible the distortions of the spring due to the radial force. The construction and method of fixing the spring provide not only its linear characteristic, but also a high natural frequency of oscillations.

The transducer signals are transmitted along conductors, placed in the central holes of the mandrels, and the slip-ring devices 1 and 18 to a special double-channel electronic amplifier 19, and then to a loop oscilloscope 20. At the same time the excessive angular displacements of the gears can be observed and measured approximately on the cathode-ray oscilloscope 21.

* The test rack was designed by M. A. Skuridin, G. B. Kozhevnikov, and L. A. Kudryavtsev.

The electron amplifier 19 is equipped with a low-pass filter which only passes cyclic errors of the gears, disciminating against kinematic errors.

For the recording of kinematic errors standard amplifiers without low-pass filters are used.

Figure 2 provides as an example a part of an oscillogram of excessive displacemtns of the driven gear (curve ΔF) obtained on the test rack. The experiment was carried out with spur gears: $z_1 = z_2 = 32$, $m = 4$ mm, $n = 630$ rpm for a relative load on the teeth of 110 kg -wt/ cm; the gears were of a 6th grade of accuracy according to GOST (All-Union State Standard) 1643-56.

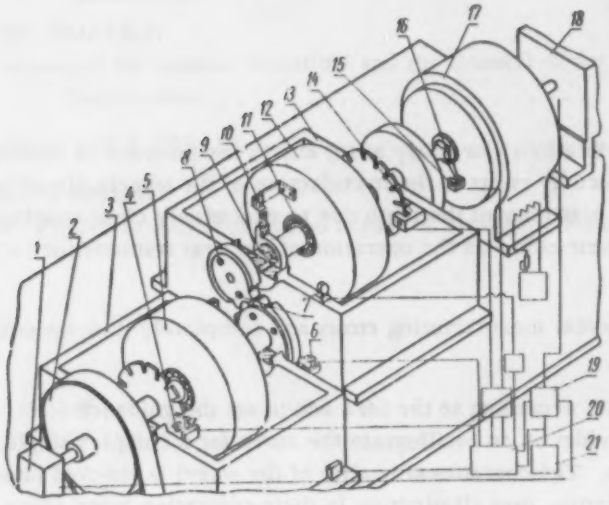

Fig. 1.

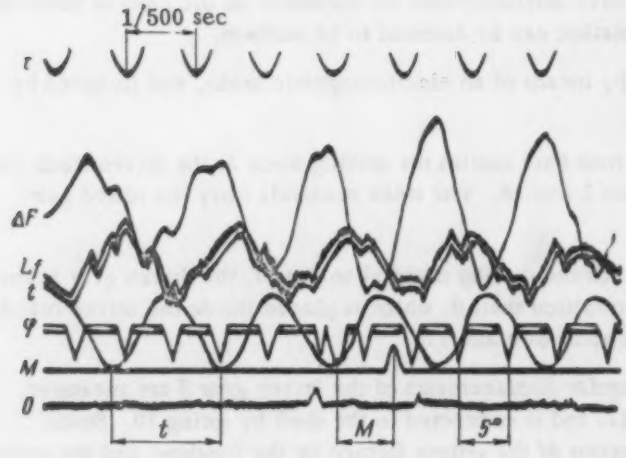

Fig. 2.

In order to ascertain to which teeth certain portions of the oscillograms refer, the test rack is fitted with a rotation angle transducer of the driven gear.

This transducer consists of a photodiode 14 (see Fig. 1), an illuminator and a disc with radial slots. It produces curve φ on the oscillogram.

The value of the braking torque is determined by the torsion shaft 9 twist angle, which is indicated by pointer 16 deflections with respect to the zero slot of the disc. The twist angle is recorded by means of a photoelectric transducer 15, which is similar to the above mentioned transducer 14. The value of the braking torque can be determined from the oscillogram by measuring the displacement of the mark on curve M with respect to the zero mark on curve φ.

For testing gears with a gear ratio of $i \neq 1$ a second induction transducer 4 is fitted for determining the angle of rotation of the driving gear.

Induction transducer 6 serves to mark on the oscillogram the instant a new pair of teeth engage, and hence to divide the oscillogram into sections corresponding to separate steps (Fig. 2, curve t).

The lack of contact between the teeth of the engaging gears is controlled by means of the braking of the electrical circuit between the insulated driving and driven stocks of the rack (Fig. 2, curve O). Curve τ provides a time scale, each period corresponding to 1/ 500 sec.

The elastic inertial parameters of the test rack can be measured by means of exchangeable discs, replaceable torsion shafts, and by connecting different masses to the tested gears.

The construction of the test rack is designed to isolated dynamically the mandrel together with the tested gears from oscillations of the rotating shafts of the motor and the brake. For this purpose flywheels with a large moment of inertia are mounted on the driving mandrel and the shaft of the brake, and the driven shaft is connected to the shaft of the brake by means of torsion shaft 9 which has a small twist rigidity.

The quality of the gear wheel operation evaluated by means of the transducer of excessive displacements is compared with the evaluation obtained from microphone MD-44 and a special electronic amplifier incorporated in the test rack. Experiments were made with simultaneous oscillographic recordings of the excessive displacement of the driving wheel (curve ΔF) and the noise (curve L_f) produced by cyclic errors (accumulated errors were outside the microphone amplifier bandwidth). The comparison of curves ΔF and L_f shows their similar nature. A certain lag in the noise curve is due to the time taken by the sound to be transmitted from the meshing zone of the gears

to the microphone. It will be seen from the oscillogram that the excessive displacements transducer provides an evaluation of the operation of each tooth. This is seen from curve ΔF which has in sections corresponding to various teeth different amplitudes. The values of these amplitudes agree with the results obtained from element-by-element checking of gears and with the single-profile static testing. The noise curve L_f has approximately the same amplitude over all its sections, and,hence, does not evaluate the operation of each tooth separately.

The correct evaluation of the gear transmission operation by means of the excessive displacements transducer is confirmed by a number of experiments which make it possible to arrive at the conclusion that the measurements made on the test rack by means of the excessive displacements transducer characterize correctly the operation of gear drives. As distinct from noise measurements, it is possible by its means to evaluate the operation of gears for each tooth, and to determine quantitatively the excess rotation angles of the gears; whereas the transmitted noise depends not only on the accuracy of the wheel manufacture, but also on their construction and other acoustical conditions.

The investigations carried out on the test rack showed that it is possible by means of the oscillograms obtained on it to find the meshing sections with an unsatisfactory operation of the teeth, and to evaluate the extent to which any manufacturing error affects the quality of gear transmission at working loads and speeds.

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INVESTIGATING THE EFFECT OF THE MICROINTERFEROMETER APERTURE ON THE ACCURACY OF MEASURING THE HEIGHT OF IRREGULARITIES IN A TRIANGULAR PROFILE

I. I. Dukhopel

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pp. 3-6, July, 1961

It is often necessary to measure the height of triangular profile microirregularities by means of a microinterferometer. Irregularities of such a type include, for instance, the lines of diffraction gratings, as well as the majority of the traces of machining on metallic surfaces. Their height $\underline{h}$ is calculated by means of the same formula [1,2] which is used for evaluating the height of steps, i. e.

$$h = \frac{N\lambda}{1+\cos u}, \tag{1}$$

where N is the value of the interference fringe distortion due to an irregularity expressed by the number of fringes; λ is the light wavelength; $\underline{u}$ is the aperture angle of the beam leaving the objective.

In the case when the aperture of the beam coincides with the relative aperture A of the objective, i. e. when $\sin u = A$, expression (1) is converted to the formula quoted by A. N. Zakhar'evskii [2]:

$$h = N\frac{\lambda}{1+\sqrt{1-A^2}}. \tag{2}$$

In practice, however, the above coincidence does not always occur. Thus, in working with microinterferometer MII-4 the contrast of the image is normally increased by reducing the size of the diaphragm aperture. In this case only part of the aperture visible at the output of the microobjective is filled with the image of the light source, and hence sin u< A. This should be taken into account in precision measurements.

Fig. 1.

Fig. 2. a) N=1.7; sin u=A=0.65; b) N=2, A=0.65, $\mu \approx 10°$; c) N=1.3, sin u = A = 0.16.

According to (1) the accuracy of measuring height h does not depend either on the width of the irregularities or on the resolution of the microinterferometer, which, however, contradicts the experimental data obtained. For instance, it was found that microinterferometer IZK-50, which was previously produced, does not provide owing to its small resolution the required accuracy in checking surface roughness. In 1950 its production was discontinued, and it was replaced by microinterferometer MII-4, which has an aperture ratio of 0.65. It is also know that it is impossible to measure the height of irregularities by means of microinterferometers whose resolving distance σ is larger than the width of the irregularity.

Not all the interferometer instruments intended for checking surface roughness can, in practice, be used with large aperture ratios. Such instruments include, for instance the microprofilometer of V. P. Linnik [3] and of Yu. V. Kolomiitsov[4,5]. In this connection it is important to find the minimum aperture ratio of a microinterferometer for which it is possible to check surface roughness without appreciable errors.

The author of this article has investigated experimentally, by means of triangular profile grooves, the relation between the readings of a microinterferometer and the aperture ratio of its objective. The objects of the investigation consisted of machining traces on 18 grade 11-14 surfaces, as well as 11 triangular profile grooves of a height of 0.1 to 1 μ specially drawn on a polished surface of a steel sample. Prior to the investigation the parameters of the scratched grooves, height h and apex angle α (profile angle) were measured. Angles α of six grooves were measured at a magnification of 6000 on an electronic microscope from the profiles of their replicas. The remaining grooves were drawn by means of cutters with known rakes. The actual values of angle α were checked by calculations from the measured values of the width and height of the scratched grooves. Among the measured values of angles α the majority were over 150°; however, some of them attained 120° and less.

All the measured grooves can be divided into two groups, those with profiles approaching the ideal, and those considerably deviating from it.

The profile of the first type of grooves (Fig. 1 a) consists of a combination of relatively long straight lines with curved elements adjacent to the apex and the base. The apexes of the above grooves are rounded off over a distance of 0.5-0.7 μ in a straight line parallel to their base.

The profile of the second type of grooves (Fig. 1 b) consists of a combination of straight lines and curves of a complex form. The size of the apex of these grooves is, as a rule more than 0.7 μ. The photographs of their profiles shown in Fig. 1 were obtained from replicas by means of an electron microscope.

The real height h_0 of the groove was assumed to be that measured on a microinterferometer MII-4 (A=0.65) for a $u \approx 10°$ and calculated from (1).

The value of the interference fringe irregularities due to the scratched grooves was measured for various apertures and magnifications of the microinterferometer.* For these measurements an interferometer was assembled

* In these measurements the condition sin u=A was observed.

TABLE 1

Number of groove	Height h_0, μ	Profile angle α, deg.	Objective aperture ratio A	Irregularity N of an interference fringe	
				Measured	Calculated from (2)
1	0.79	120	0.65	2.6	2.54
			0.43	2.6	2.72
			0.33	1.8	2.81
			0.22	1.6	2.86
			0.16	1.1	2.88
			0.11	1.0	2.90
2	0.74	132	0.65	2.4	2.38
			0.43	2.5	2.54
			0.33	2.1	2.63
			0.22	1.5	2.67
			0.16	1.2	2.69
			0.11	0.9	2.70
3	0.52	150	0.65	1.7	1.67
			0.43	1.8	1.82
			0.33	1.8	1.85
			0.22	1.9	1.88
			0.16	1.7	1.89
			0.11	1.6	1.90
4	0.3	*	0.65	1.0	0.97
			0.43	1.0	1.04
			0.33	1.1	1.07
			0.22	1.1	1.08
			0.16	1.0	1.09
			0.11	0.7	1.10

TABLE 2

Grade of surface finish	Irregularity N of an interference fringe for aperture ratio A				
	0.65	0.43	0.33	0.22	0.16
11	1.5	1.6	1.7	0.9	0.7
12	0.9	1.0	1.0	0.8	-
13	0.6	0.7	0.7	0.4	-

which was similar to V. P. Linnik's double-objective interferometer attachment [6], which fits over the microscope tube instead of the objective. The simplicity of this device made it easy to measure the aperture and magnification of the equipment.

In order to check the irregularities observed in the interference fringes due to the scratched grooves, they were measured by means of another two single-objective interferometer attachments, whose aperture ratios were 0.25 and 0.17 respectively. The source of light in these measurements consisted of a mercury-arc microlamp (λ=5461A).

The measurement results of four grooves are given in Table 1.* The profiles of the first two grooves in the table are of the first type. The last two age of the second type. The interference fringe irregularity measured in grooves 3 and 4 represents the arithmetic mean of measurements made at magnifications of 800, 600, 400, 280 and 160, and in grooves 1 and 2 the arithmetic mean of measurements made at magnifications of 800, 600, and 400. The readings obtained in grooves 1 and 2 at magnifications of 280 and 160 were rejected as unreliable. Side by side with the measured values of N the table contains the respective irregularities of interference fringes calculated from (2). These measurements have shown the following.

Firstly, in grooves with a profile approaching the ideal (see Table 1, grooves 1 and 2) the value of the irregularity of interference fringes decreased with a decreasing microinterferometer aperture ratio, becoming more pronounced with a smaller profile angle α of the groove. Irregularities observed with aperture ratios exceeding 0.40 formed an exception. Their values agreed well with those calculated from (2).

Secondly, in grooves of the second type (see Table 1, grooves 3 and 4) the value of the irregularities in interference fringes increased with the aperture ratio decreasing to 0.22 or less, and the values of N measured and calculated from (2) approached each other. A further reduction in the aperture led to a decreasing irregularity in the fringes.

Figure 2 gives a graphic representation of the variation in the irregularity of interference fringes with changes in u and A. It shows three interference pattern photographs of a groove with h=0.54 μ and α=160°.

* The nature of the variations in the shape of the remaining seven grooves is the same, and, hence, the results of their measurements are not given.

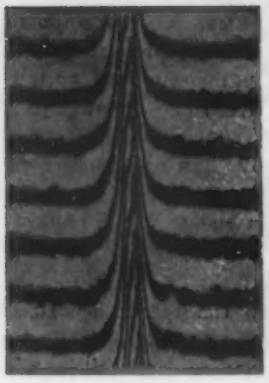

Fig. 3.

In grooves with a large value of h one can observe in addition to the gradual variation in the value of the interference fringe irregularity, also a sudden change which it is difficult to understand. This occurs more often in grooves with a small profile angle α.

The theoretical explanation of the interference patterns of triangular profile grooves is a very complicated problem which has not yet been solved. We can only assume that the size of the interference fringe irregularity observed by means of the microinterferometer depends on the relation between the dimensions of the groove apex and the resolving distance σ of the instrument. It is probable that when the size of the apex is larger than σ, the groove may be regarded as a step and its height calculated without appreciable error from (2).

Investigations of the traces of machining on metallic surfaces led to results similar to those observed for grooves 3 and 4 (see Table 1). In 94% of the observed grooves a decrease of the microinterferometer aperture led to a reduction in the interference fringe irregularity; moreover, the value of the irregularity calculated from (2) agreed well with the measured value. A further reduction in the aperture led to a decrease in irregularity N. As an example Table 2 provides measurement results for three grooves on a surface with a grade 11-13 finish.

In evaluating the irregularity of fringes an important part is played by the microinterferometer magnification in addition to its aperture ratio. We have previously discarded measurements of grooves with a large height and a small profile angle α as unreliable with a magnification below 300. The interference fringes at the edges of such grooves are placed so close to each other that for a small microinterferometer magnification it becomes difficult or impossible to measure the value of the irregularity in one of them. As an example Fig. 3 shows an interference pattern photograph of a groove with slope angles of its side* equal to 24° and 45°. The microinterferometer magnification was 490, its aperture ratio 0.65, and $u \approx 10°$. It will be seen from Fig. 3 that a reading of the irregularity which equals four fringes can only be taken on the left-hand side, whose angle of slope is equal to 24°.

It has been established experimentally that a reliable evaluation of the value of an interference fringe irregularity can be made only when the width of the fringe visible at the side of the groove amounts to at least 0.3-0.4 mm. On machined surfaces this condition is met with a microinterferometer magnification of the order of 300-400.

CONCLUSIONS

1. In order to obtain correct information on the height of a triangular profile groove mearurements should be made by means of a microinterferometer whose resolving distance σ should not exceed the width of the groove.

2. For a correct determination of the finished grade of machined surfaces it is necessary to use microinterferometers whose objectives have an aperture ratio of 0.30.

3. Microinterferometers used for checking surface roughness should have a magnification of not less than 300.

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* The angle formed by the side of the irregularity with the basic surface plane.

MEASURING THE PROFILE ANGLE OF INTERNAL CONICAL THREADS

A. A. Khairov

Translated from Izmeritel'naya Tekhnika, No. 7,
pp. 6-7, July, 1961

In cutting and repairing the threads of boring lock couplings the profile angle of the thread has hitherto not been measured.

The author of this article has developed and made in the area laboratory of the "Tuimazaburneft" trust a portable electrical contact instrument for measuring half the profile angle of internal threads with a pitch of 6 and 6.35 mm. With an appropriate change in the dimension of the electrical contact head it is possible to measure profiles of threads with a pitch of 5 mm.

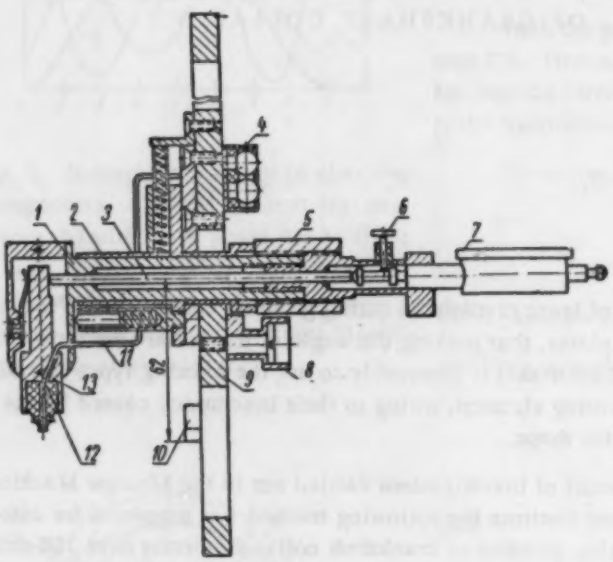


Fig. 1.

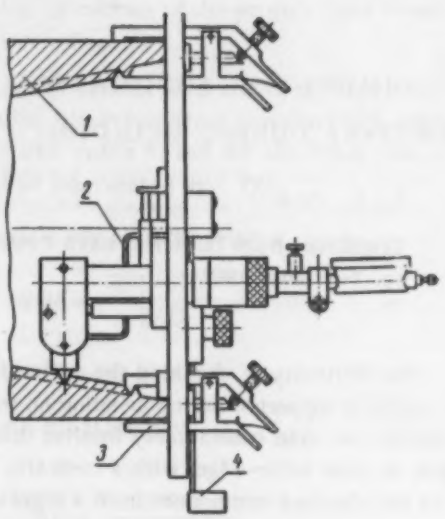


Fig. 2. Instrument in the operating position. 1) The boring lock coupling; 2) electrical cable to the contact head; 3) clamps; 4) power pack.

The instrument consists (Fig. 1) of bar 9 over which cantilever bracket 2 moves perpendicularly to the axis of the coupling and carries inside it rod 1 which is displaced along the axis. The front end of rod 1 carries an electrical contact template 13 which swings about axis 12. The rear end of rod 1 is connected to indicating gauge 7. The upper end of the template is connected to the gauge by means of rod 3. The lower end of the template is fitted with an electrical contact head secured by two screws. The head has four measuring insulated contacts connected by means of cable 11 to the electrical bulbs of the display board 4. The current is supplied by a 1.5 v battery type BAS-8. Spark-quenching capacitors are used across the contacts.

Measurements are started by turning catch 8, placing bracket 2 into its upper position, inserting the instrument into the measured article, resting it against the centering stops 10 and pressing bar 9 to the butt end of the article. The instrument is then secured by means of the clamps (see Fig. 2, 3). The instrument is adjusted in such a manner that in this position the electrical contact template 13 will be placed in a position perpendicular to the axis of the article. By displacing rod 1 by means of nut 5 along the axis, and at the same time gradually lowering bracket 2 by means of catch 8, the contact head is inserted into one of the thread grooves. By observing the lighting of the two lower bulbs of display board 4 it is possible to ascertain whether the lower contacts (projections) of template 13 head are touching evenly both sides of the thread groove. This position of index pin 6 corresponds to the initial (zero) reading of the indicator gauge. Nut 5 is then turned clockwise and the instant at which the right-hand side bulb of display board 4 lights is observed, thus indicating whether the top right-hand contact is touching the profile of the

thread. As soon as the bulb lights the turning of nut 5 is discontinued and the reading of indicator gauge 7 noted. By turning nut 5 counterclockwise the instant at which the left-hand bulb lights is observed, and the reading of the indicator gauge is again noted. The first and second readings of the indicator correspond to the actual value of half the thread profile angle.

Axis 12 of template 13 is placed with respect to the axis of rod 3 in such a manner that one indicator division corresponds to one minute. The measuring range of half the profile angle amounts to 30° ± 1°.

The error in the instrument readings for a 6th grade surface finish of the thread amounts to 6'. It takes 3 min to measure one article.

It is possible by means of this instrument to measure thread profile angles in situ. Its weight amounts to 1.2 kg, and that of the battery in its casing to 0.8 kg.

INSTRUMENT FOR AUTOMATIC TESTING OF CRANKSHAFT COLLAR DIAMETERS DURING GRINDING

E. I. Ped'

Translated from Izmeritel'naya Tekhnika, No. 7,
pp. 7-8, July, 1961

The difficulty in checking the collar diameters of large crankshafts during grinding consists in the fact that each collar is supported during grinding on two collar plates, thus making the angle of the collar which the measuring instrument can span considerably smaller than 180°. This makes it impossible to use the existing types of caliper gauges, or rider knife-edges with a centrally placed sensing element, owing to their inaccuracy caused by the deviation of the checked component from a regular geometric shape.

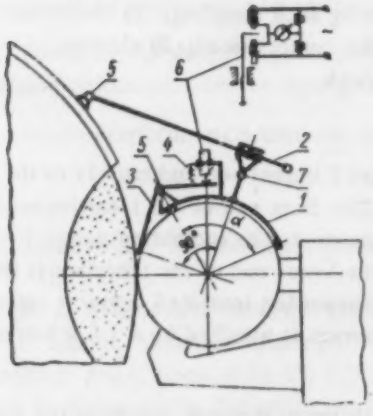


Fig. 1.

As a result of investigations carried out in the Moscow Machine-Tool and Instrument Institute the following method was suggested for automatic checking during grinding of crankshaft collar diameters over 150 mm with a tolerance of 20 μ.

The schematic of this instrument is shown in Fig. 1.

The shaft collar diameter is measured during grinding in the following manner: Measuring prism 1 is placed on the machined detail by turning lever 2. The prism rests on the shaft on two hard-alloy cylindrical bearings, and a third mobile bearing is fitted at the end of measuring lever 3. Spring 4 provides the measuring effort of the instrument. With a reduction in the size allowed for grinding measuring lever 3 turns about its hinge 5 and displaces with its other arm the measuring rod of induction transducer 6, which is connected to a bridge circuit and continuously checks the size of the machined component's radius.

The self-alignment of the prism on the shaft is provided by hinges 5 and 7 which have no backlash and prevent any wobble in the shaft affecting the measurement results.

The instrument is set by means of a sample component.

By comparing the effect of deviations of the measured component from a regular geometrical shape on the measurement results, with similar values obtained in checking by means of rider prisms, which are now used in industry, it becomes possible to assert that the above arrangement has certain advantages.

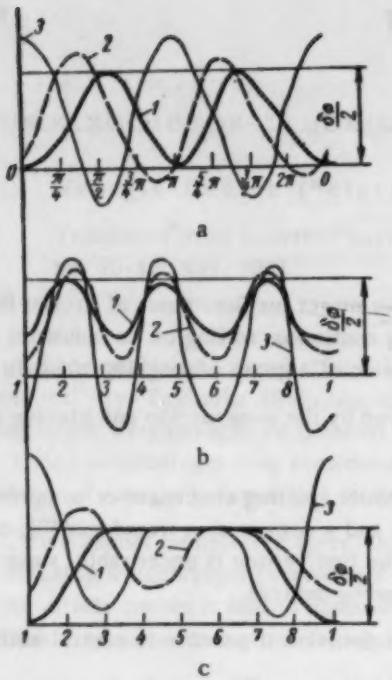


Fig. 2. Instrument readings in checking components. a) with a symmetrical oval shape; b) with three peaks due to flats; c) with an asymmetrical oval shape; 1) actual deviations of the component's radius; 2) instruments readings; 3) readings of the rider prisms.

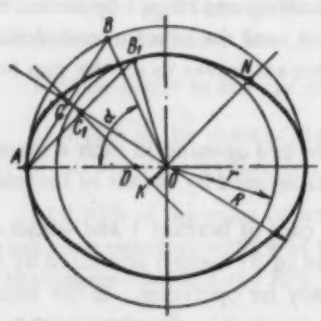


Fig. 3.

For comparison purposes Fig. 2 provides readings of the above instrument with angle $\alpha=90°$ and those of rider prisms with a centrally placed sensing element and angle $\alpha=120°$.

Thus, in both arrangements the rods of the sensing elements (transducers) will be displaced by an amount $\delta/2$ for a deviation of the checked component's radius by δ.

The maximum error due to the component's deviation from a regular geometrical shape can be calculated from the drawing shown in Fig. 3. Let us assume that

$$R-r=\frac{\delta\varnothing}{2},$$

where $\delta\Phi$ is the allowance for a deviation from a regular geometrical shape; A and B are the points of contact of the prism's fixed bearings; α is the central angle between them.

With the prism's points of contact at A and B_1 the instrument measures KN. Hence, for the most disadvantageous position when one bearing touches the circumference with radius R, and the other that with radius r, the maximum error will be represented by ± OK.

By solving triangle AC_1D and DKO we have

$$\pm OK=\frac{b(R-AD)}{2AD},$$

where $b=AB=2R\sin\frac{\alpha}{2}$;

and from triangles AB_1O and AC_1D we find

$$AD=\frac{b^2R}{b^2+R^2-r^2}.$$

then

$$\pm OK=\pm\varepsilon\varnothing=\frac{(R+r)(R-r)}{4R\sin\frac{\alpha}{2}},$$

where

$$R-r=\frac{\delta\varnothing}{2}.$$

Hence

$$\pm\varepsilon\varnothing=\frac{\delta\varnothing(R+r)}{8R\sin\frac{\alpha}{2}};$$

α should be between 90 and 120°.

Thus, if the value of the error due to deviations of the component from a regular geometrical shape is known, it is necessary to take this into account in setting the transducer.

The above considerations were confirmed by the use of the instruments in production, when they measured on grinding machine multiple crankshaft collars 200± 0.015 mm in diameter.

INSTRUMENT FOR MEASURING STRESSES IN STRIPS OF CONTINUOUS SECTION MILLS

V. N. Vydrin and P. N. Amosov

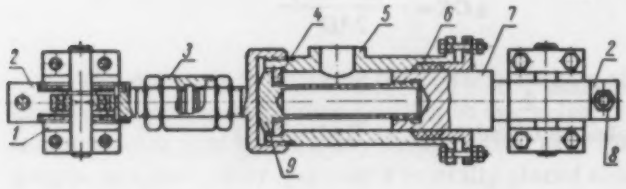
Translated from Izmeritel'naya Tekhnika, No. 7,
pp. 9, July, 1961

In 1956 the Chelyabinsk Polytechnical Institute suggested a method for direct measurements of stresses in continuous section mills [1,2], which was successfully used for investigating continuous rolling on an industrial mill. At present it is intended to use this method in controlling high-speed operation of a group of roughing stands in a mill.

An important part in the successful application of this method is played by the construction and placing of the stress transducer.

In the continuous operation of a mill the strip may be subjected to tensile (pulling condition) or compression (backing) stresses. Hence, the stand must be equipped either with a tensile and a compression transducer [2], or with a single transducer which registers both tensile and compression stresses. The first version is undesirable, since it requires a double quantity of transducers, expensive screen leads, and measuring devices.

For this purpose a new tensile transducer was developed (see Fig.) which makes it possible to control both the pulling and backing of the stand on both sides.



The transducer casing 5 contains a core 9 which carries a strain gauge bridge. The core is fixed without backlash to the transducer casing by means of ring 4, a lid with an internal thread and rod 7 with an internal thread. The pipe thread in the lid and gland 6 protect the transducer cavity from moisture and dust. The two bearing housings 2 hold double-row ball-bearings No. 1002 which are secured by means of clamping bolts 8. The bearings are mounted on pins and protected from moisture by rubber packing and rings. By means of caps and bolts one of the bearing pins is fixed to brackets 1 which are welded to the bed, and the other to connecting lugs welded to the roller blocks. Adjusting nut 3 with left-hand and right-hand threads serves to set the gap between the bed and the block.

By its means the block, which rests on a cylindrical bearing, is placed in the bed opening in such a manner that equal gaps are formed on either of its sides. In this position adjusting nut 3 is secured by means of locknuts.

In changing the rollers it is only necessary to loosen four bolts, lift the two caps of bracket 1 and withdraw the transducer. Bracket 1 must already be welded to the new block. The pin is placed in its sockets and fixed by means of the caps. The required gap is fixed by means of nut 3 and the transducer is ready for operation. In the adjustment of the rollers or the replacement of blocks the fixing points of the transducer may be displaced with respect to each other. Additional bending stresses in the transducer in such a case are eliminated by the self-adjusting double-row ball-bearings.

The above design of the tensile transducer makes it possible to measure both the pull and the backing on either side of the stand; to determine the effect of the pulling and backing on the rolling pressure, the forward flow, shrinkage of the profile and other factors; to determine the additional horizontal loading on the roller bearings which is especially important for high-speed mills; to control the operation of the mill and automatically control the given stress by feeding the transducer pulses to an electrical circuit which controls various displacements in the mill; to provide "endless" hot section rolling with a tensioned strip, thus ensuring all the advantages of such an operation.

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2. V. N. Vydrin, Dynamics of Rolling Mills [in Russian] Metallurgizdat (Moscow, 1960).

MEASUREMENTS OF MASS

CHECKING MOTOR-TRUCK WEIGHBRIDGES

Georgie Dedich (Belgrade, F.P.R.Y.)

Translated from Izmeritel'naya Tekhnika, No. 7,
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The conclusion arrived at by S. I. Gauzner in his article on "Checking Technique of Motor-Truck Weighbridges" (Izmeritel'naya Tekhnika 1960, No. 3) that the method specified by instruction 15-52 is unsuitable for checking motor-truck weighbridges we consider to be correct. The sum of the errors in determining the mass of weights on the tested weighbridges may considerably exceed the tolerances of these bridges; moreover, in the overwhelming majority of cases it is impossible to find on the spot a load for substituting standard weights.

We cannot completely agree, however, with the second conclusion arrived at in the article, that the method of checking weighbridges by adding a small weight is unsuitable. The manner of applying this method described in the article makes it indeed unsuitable, since the authors of this method have not taken into account the errors of the beam, which were determined when it was checked on the test rack.

We suggest a modification of this method which would take into account in checking the weighbridges the errors of the beam which are determined when the bridge is checked on a test rack, thus making it possible to check motor-truck and railroad car weighbridges with sufficient accuracy.

The essence of this method consists of the following:

The beam is checked on the test rack according to instruction 15-52 with the following assumptions.

1. The beam sensitivity is tested by means of a sliding weight placed on the last scale calibration and on the calibration corresponding to 1/5 of the scale.

The sensitivity of the beam is considered satisfactory if the difference in the beam deflection between the first and second positions of the weight is at least 5 mm, the weight being equal to 0.02% of the total mass of weights which balance the beam for a maximum load.

2. The errors in the cut grooves should not exceed:

a) 0.05% of the mass of standard weights which are placed in the pan for checking each calibration, in the interval between the calibration corresponding to the total load and the calibration corresponding to 20% of it;

b) 0.05% of the mass of standard weights which balance the beam with the sliding weight placed on the calibration corresponding to 20% of the maximum load, in the interval between the calibration corresponding to 20% of the maximum load and the first significant calibration.

3. In checking the auxiliary scale the error should not exceed 0.05% of the mass of standard weights which balance the bean with the sliding weight placed on the calibration corresponding to 10% of the maximum load.

For any further on-the-spot checking of the weighbridge according to the suggested method it is necessary to determine the errors of the beam with the sliding weight placed on the calibration K_1 corresponding to 20% of the maximum scale reading, and on the calibration K_2 corresponding to 80% of that reading. These errors should be entered in the certificate of the beam.

In order to explain the method for checking the errors of the cut grooves on the main scale we shall give the following example.

The weighbridge scale up to 10 tons has 20 grooves (without counting the zero grove). In order to balance it with a sliding weight placed on the last calibration of the scale 21 kg had to be placed in the pan, and hence, the value of one scale division is equal to 21÷20=1.05 kg.

The mass of standard weights for the 4th calibration of the scale (20% of the scale) should be equal to $1.05 \times 4 = 4.2$ kg. In checking we obtained 4.198 kg. The error $K_1 = 4.2 - 4.198 = 0.002$ kg.

Thus, the tolerance is equal to $\pm\frac{4200 \cdot 0.05}{100} = \pm\ 2.1$ g.

K_2 is determined similarly for the 16th calibration of the scale.

By means of a beam certificate with the values of K_1 and K_2 entered on it, and knowing that the error of the zero and maximum scale readings is equal to nought, the inspector can check the weighbridge on the spot by the method of an additional small weight. This test consists of the following.

The weighbridge is balanced by means of the tare weight regulator and then deliberately pushed off balance. It should return to balance either automatically or by placing on the weighbridge platform a weight not exceeding 0.01% of the maximum load. A motor-truck loaded up to 60-80% of the bridge capacity should then be driven at a rate not exceeding 5 km /hr across the weighbridge, first from on and then from the other side. The difference in the readings of the bridge should then not exceed 0.01% of the maximum load.

If the bridge has an isolator, the tests for a constant balance should be repeated three times with the isolator closed and opened for each balance.

The weighbridges which satisfy the above requirements are again balanced and checked by means of standard weights. For this purpose standard weights of a mass of 20% of the maximum bridge load are placed consecutively on the platform of the bridge over each pair of knife-edge bearings with the detainer first closed and then opened for checking the bridge readings. If the weighbridge cannot be balanced, a tolerance weight equal to 0.02% of the maximum load is placed on the platform (or removed), in order to obtain a balance or swing the beam in the opposite direction. In this operation the displacement of the balance indicator must not be less than 2 mm (sensitivity check).

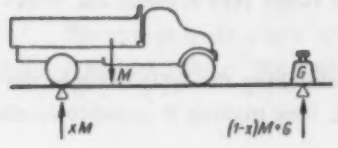

Fig. 1.

The difference in the bridge readings when checking at various points of the platform must not exceed 0.02% of the maximum load.

The standard weights should then be placed in the middle of the platform and balanced by the sliding weight in the appropriate position. If the bridge becomes unbalanced it should be rebalanced by adding where necessary a tolerance load of $T_1 = K_1 + P_1$, where K_1 is the error obtained in checking the beam on the test rack, and P_1 is the error in the lever ratio.

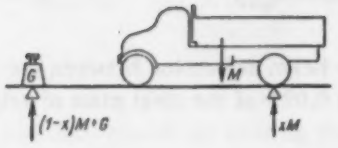

Fig. 2.

Error P_1 must not exceed 0.01%, and error $T_1 - 0.02\%$ of the maximum load.

Having removed the standard weights from the platform, the truck is driven onto it and the sliding weight set to the calibration corresponding to 80% of the maximum scale reading, and the weighbridge brought to balance by adding a load (ballast) onto the truck. The sliding weight is then placed on the maximum calibration, and standard weights of a mass of 20% of the maximum load are placed onto the platform. Should the bridge then become unbalanced, the balance should be restored by adding in the appropriate place weight T_2 whose mass should not exceed 0.05% of the maximum load.

The error of the lever ratio is equal to $P_2 = T_2 - K_2$. Having determined P_2 the errors of P_1 and P_2 are compared.

If the difference between P_1 and P_2 does not exceed 0.01% of the maximum load the weighbridge is passed for use.

In evaluating the method for checking the scales suggested by I. S. Krever, it should be noted that the above method cannot always be applied, since in certain types of beams it is impossible to place the apexes of the carriage knife-edge bearings in line with those of the beam bearings, and also because some of the earlier weighbridge models do not have sufficient room for fitting the carriage.

In spite of this the above method is more convenient than checking by means of a ballast load, and it should be used with a preliminary verification made by the above-mentioned method.

The checking of the beam on the test rack, determining the reading stability of an unloaded weighbridge, evaluating the deviation in readings when a truck loaded up to 80-90% of the bridge capacity is driven across it, and checking the bridge at 10% of its maximum load is carried out according to the technique described in this

article, but the tolerance in this case must not exceed 0.01% of the maximum load.

Having determined the above errors the truck is weighed twice by driving it from two different sides onto the platform. The difference in the two readings must not exceed 0.05% of the maximum load. At the same time the sensitivity of the bridge is tested by changing the load by 0.02% of its maximum; the beam zero indicator should then be displaced by at least 2 mm.

The ratio P_0 of the bridge levers is then determined. For this purpose standard grade 3 weights totalling G=10% of the maximum load are placed on the middle of the platform and the weighbridge is balanced by grade 2 weights placed on the carriage pan. The lever ratio will be

$$P_0 = \frac{A_0}{G}, \tag{1}$$

where A_0 is the mass of the grade 2 standard weights in the pan.

After balance has been obtained the standard weights are removed from the platform and the truck is driven onto it in such a manner as to place its rear wheels over the left-hand knife-edge bearings (Fig. 1), and is weighed by placing on the carriage pan grade 2 weights of a mass A_1. As a result of this we obtain the equation

$$xMP_1+(1-x)\,MP_2=A_1. \tag{2}$$

Standard grade 3 weights of a mass G = 10% of the maximum load are then placed over the right-hand knife-edge bearings and the bridge is balanced by placing in the carriage pan standard grade 2 weights of a mass B_1. The corresponding equation will be

$$xMP_1+(1-x)\,MP_2+GP_2=B_1. \tag{3}$$

Similar measurements are then made with the rear wheels of the truck placed over the right-hand bearings of the weigh-bridge (Fig. 2). We obtain the equations

$$(1-x)\,MP_1+xMP_2=A_2, \tag{4}$$

$$(1-x)\,MP_1+xMP_2+GP_1=B_2. \tag{5}$$

Grade 3 standard weights of a mass G = 10% of the total load are then again placed in the middle of the platform and the bridge is balanced by placing in the carriage pan standard grade 2 weights of a mass A_2. The ratio of the lever arms is then determined for an unloaded weighbridge as:

$$P_2 = \frac{A_2}{G}. \tag{6}$$

The mass of the truck is then determined.

From (2) and (4) we obtain the computed mass of the truck:

$$M = \frac{A_1+A_2}{P_1+P_2}. \tag{7}$$

We obtain the ratio of the lever arms from (3) and (5).

$$P_2 = \frac{B_1-A_1}{G}, \tag{8}$$

$$P_1 = \frac{B_2-A_2}{G}. \tag{9}$$

The mean ratio of the lever arms

$$P=\frac{B_1+B_2-A_1-A_2}{2G}. \tag{10}$$

and

$$\frac{P_0+P_3}{2}=\frac{A_0+A_3}{2G} \tag{11}$$

should not differ from the above values by more than $0.002\ (P_0+P_3)/2$, i. e.

$$B_1+B_2-A_1-A_2-A_0-A_3<\pm 0.002\,(A_0+A_3). \tag{12}$$

The difference between P_0 and P_3 should not exceed

$$0.002\ \frac{P_0+P_3}{2}.$$

Having determined the computed mass M of the truck, the standard weights are taken off the bridge platform and the carriage pan, and the actual mass M_a of the truck is determined by means of the sliding weight.

If the difference between M_a and M does not exceed 0.1% of M, the weighbridge is considered satisfactory for use.

MECHANICAL MEASUREMENTS

EXPERIENCE IN PRODUCING A REFERENCE ACCELEROMETER

P. N. Agaletskii, V. Ya. Barash, S. A. Bogdanova,

and Zh. P. Nikulina

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The basic characteristics of an accelerometer and their agreement with the specifications are determined in testing the particular accelerometer type; the instrument is checked for correct and accurate readings within its frequency range when it is calibrated or checked. For accelerometers whose readings are expressed in conventional units (in scale divisions or millivolts) the concepts of calibration and checking are essentially the same.

For linear scale accelerometers whose readings are expressed in conventional units the instrument constant subject to inspection and checking is taken to be the calibration coefficient determined as the mean value of sensitivity in the operating frequency range. Normally the dynamic calibration of an accelerometer is made for a given frequency and amplitude of sinusoidal vibrations impressed on the instrument by means of an appropriate vibrator.

For calibration purposes sinusoidal vibration generators are normally used with means for measuring the acceleration induced in the calibrated accelerometer. This acceleration is determined indirectly from direct measurements of amplitude A and frequency f.

The error of indirect measurements depends not only on the errors in measuring A and f, but also on the extent to which the quantities whose values are not measured but affect the results are studied and accounted for.

The results of indirect measurements carried out according to the same method but with different apparatus and under different conditions may contain different systematic errors depending on the extent to which the quantities affecting the measurement results have been isolated, studied and accounted for.

It is an established practice in dynamic calibrations of accelerometers to calculate the acceleration amplitude a from the formula $a = 4\pi^2 f^2 A$ even when the waveform of the vibration test rack platform oscillations is not sinusoidal. Owing to this fact, the instability of the vibration frequency and the low accuracy of its measurement, the error in calibrating accelerometers under production conditions by measuring the amplitude and frequency of the test-rack oscillations amounts to some 10%. Only when all the sources of errors are carefully investigated, the amplitude and frequency measured with high precision, and special vibrators used, is it possible to reduce this error to 1/5 or 1/10 of its former value.

The calibration of accelerometers by indirect measurements of the vibration test-rack acceleration is, moreover, an inefficient operation.

The calibration is simplified and the vibrator requirements made considerably easier if the acceleration transmitted to the calibrated accelerometer instead of being measured indirectly is measured directly by means of a reference accelerometer. The readings of the tested and reference accelerometer are then taken simultaneously as the instruments are subjected to the same acceleration. Then the exciter is only required to provide the tested and reference accelerometer with the same accelerations in a given frequency and amplitude range. If the two compared instruments are linear, placed coaxially on the vibration rack and firmly fixed, and we denote the output signal and sensitivity of the reference accelerometer by x_1 and S_1 respectively and that of the tested accelerometer by x_2 and S_2, we shall obtain the following relation

$$S_2 = \frac{x_2}{x_1} S_1.$$

However, this calibration method has not been adopted owing to the lack of reference accelerometers. In order to be able to apply in practice the comparison method of testing accelerometers it is necessary to specify the

requirements of the reference accelerometer for various measuring and frequency ranges, develop typical designs for reference accelerometers which could be recommended to instrument-making plants for their production in the required quantities, to develop the methods and means for certifying reference accelerometers in metrological organizations, and for inspection testing of operating accelerometers, to specify typical designs for exciters intended for calibration circuits, and to organize their mass production.

The experience gained by the VNIIK (All-Union Scientific Research Institute of the Committee of Standards, Measures and Measuring Instruments) dynamic measurements laboratory in solving the above-mentioned problems is given below.

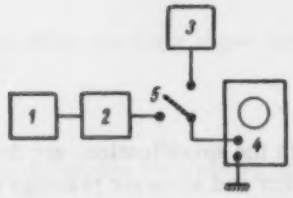

Fig. 1. 1) Piezoelectric accelerometer transducer; 2) cathode follower; 3) ac voltage calibrator; 4) cathode-ray oscilloscope used for comparing the voltage amplitudes at the output of the cathode follower with those of the calibrator; 5) switch for the meansured and reference voltages.

It is possible to use as reference accelerometers instruments of the same construction as those being tested, providing they have a considerably smaller error as compared with the latter instruments. This can be attained if highly sensitive and stable instruments are selected and carefully calibrated by means of one of the indirect methods. However, in practice it is not advisable to have as many different types of reference accelerometers as there are commercial models, since the latter differ by their intended application, whereas the reference accelerometers are only intended to measure the acceleration transmitted by the exciter to the instrument under calibration.

We think it is possible to specify for a reference accelerometer general requirements of the following nature: stable readings (both short and long term); high precision in measuring the output signal; equal relative errors over the whole measurement range; constant sensitivity over a wide frequency range; a linear scale; and a small mass and dimensions of the transducer.

A comparative analysis of different types of accelerometers has shown that the above requirements are most fully met by a accelerometer (Fig. 1) which consists of a piezoelectric accelerometer transducer 1, cathode follower 2, and a device for measuring the peak values of the voltage at the cathode follower output. The calibration of such an accelerometer amounts to determining the ratio of the peak voltage at the cathode follower output to the peak value of the vibration impressed on the transducer.

In order to obtain over the whole range measured by the accelerometer equal relative errors, a varying sensitivity comparator was used, which consisted of a cathode-ray oscilloscope 4. The oscilloscope serves to compare the peak voltage at the cathode follower output to known voltages obtained from calibrator 3 which transmits peak voltages of E_1, E_2 and E_3, thus providing a set of equally accurate voltage values. By adjusting the amplification of the signals being compared on the oscilloscope it is possible to measure both small and large voltages with a relative error not exceeding 1.5%.

The constants of the above instrument consist of the calibration coefficient $\underline{k}$ expressed in mv/g and the peak voltages E_1, E_2, E_3 ... at the output of the voltage calibrator expressed in mv.

Accelerations are measured by means of this instrument in the following manner. The peak voltage $\underline{e}$ is first measured in millivolts at the output of the cathode follower. For this purpose the measured voltage is fed to the oscilloscope and its gain is adjusted in such a manner that the extreme points of the image are not more than 10 mm from the edge of the screen. The voltage swing M is measured in millimeters on the oscilloscope scale, the input of the oscilloscope is then switched without changing its gain to the calibrator and a voltage E_l is selected, whose value approaches as closely as possible to the measured voltage. If we denote this voltage swing by L we obtain:

$$e = \frac{M}{L} E_l.$$

The actual value of the measured acceleration amplitude is $a = e/k$ (in units of g).

An original design is not an indispensible requirement of a reference accelerometer, and therefore it is advisable in designing it to use to the maximum standard-type instruments as well as units and components of commercial

accelerometers which have proved satisfactory in operation. Thus, it is possible to use for piezoelectric accelerometers the best piezoelectric transducers of commercial accelerometers designed for the required measurement and frequency ranges.

In the reference accelerometer developed by us we used a piezoelectric transducer of one of the commercial accelerometers and a standard-type oscilloscope ÉO-7.

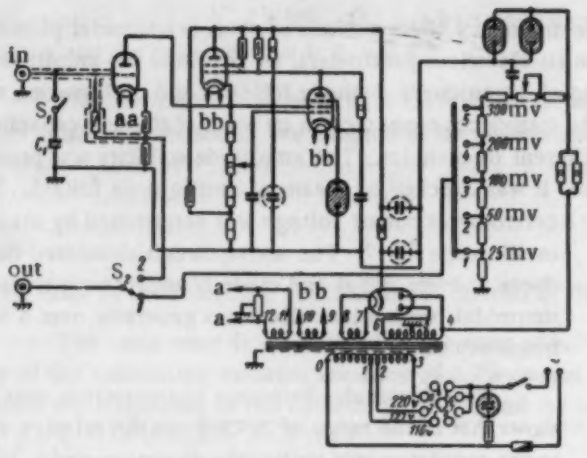

Fig. 2.

The schematic of the cathode follower and voltage calibrator which we designed for the accelerometer is shown in Fig. 2. It is known that in piezoelectric accelerometer transducers with a natural frequency of ω_0 and a damping factor of $\xi = 0.07$, the amplitude-frequency characteristic is highly distorted at frequencies approaching zero and the resonant frequency of the transducer. The distortion of the characteristics at low frequencies is due to the insufficiently large time constant of the amplifier input circuits, and at high frequencies to resonance in the mechanical oscillating system of the transducer.

The rise in the amplitude-frequency characteristic over a frequency range of $\omega = 0.25\ \omega_0$ does not exceed + 6%. Hence, in designing piezoelectric accelerometer transducers it is possible to select a value for ω_0 which would provide in a given operating frequency range a dynamic error not exceeding 1%.

At present there does not exist a standard type of piezoelectric transducer which could be recommended for a reference accelerometer. This task requires special investigation.

The production of wide-range high-precision measuring instruments is difficult, and normally separate reference instruments are used for each range.

The reference accelerometer we have developed is designed for a measuring range of 1-100 g and a frequency range of 5-100 cps. Its basic error due to the reading error, instability of readings, and calibration error amounts to ± (2-3%), and its dynamic error due to the variations of its sensitivity in the operating frequency range amounts to ± 1%. The reference accelerometer transducer weighs 30 g, its piezoelectric element made of barium titanate is 200 mm in diameter and 2 mm thick with a natural frequency of 32.5 kc and a capacitance of 1900 $\mu\mu$f. It is connected by means of cable type AVK-3. During measurements the relative variations of the input circuit capacity did not exceed ± 0.75%.

The cathode follower is mounted in the same unit as the voltage calibrator. The cathode follower supply voltage is stabilized. Its input impedance amounts to 12 meg. The instrument has two measurement ranges of 1-25 and 5-100 g. The higher range is obtained by switching in an additional capacitance C_1 in parallel with the cathode follower input. Switch S_2 serves to feed to the oscilloscope in turn the calibrating and measured voltages.

The potential divider of the calibrator is fed with a stabilized voltage in such a manner that it provides peak voltages of 25, 50, 100, 200 and 300 mv. The divider resistors are selected for an accuracy of 0.5%.

The production of a reference accelerometer is completed by its attestation. This operation consists in determining its constants and evaluating its accuracy.

For the constant of the accelerometer we used its calibration coefficient k, i. e. its mean sensitivity over the operating frequency range.

The value of k can be determined by two methods.

1. The certified instrument is fed at various frequencies within the frequency range with a sinusoidal vibration, its acceleration amplitude is measured by an indirect method, and the output voltage amplitude by a direct method, and then its sensitivity S_i is determined. The mean value of S is then taken as the calibration coefficient. The

curve of S_i versus f_i representing the relation of sensitivity to frequency is also known as the amplitude-frequency characteristic of the instrument.

2. The amplitude-frequency characteristic of the instrument is plotted in arbitrary units, and only at one frequency within the range is the sensitivity determined in appropriate units.

From this value of sensitivity its values at other frequencies within the range are determined from the amplitude-frequency characteristic, and their mean value is taken as the calibration coefficient. We used the second method since it provides by simple technical means a high accuracy in determining k.

In view of the fact that the drop in the amplitude-frequency characteristic of an accelerometer piezoelectric transducer at frequencies below $0.05\,\omega_0$ only depends on its electrical parameters, we obtained the amplitude-frequency characteristic of the electrical circuit consisting of a transducer, cathode follower and oscilloscope in the following manner. The cathode follower input (with the transducer connected to it) was fed through capacitor POV (of $390\,\mu\mu f$) by a voltage of a constant amplitude at different frequencies. The amplitude stability was provided by a regulating device incorporated in the generator, and it was checked by means of oscilloscope ÉNO-1. The accelerometer output voltage was determined by means of oscilloscope ÉO-7. The above circuit simulated the transducer's output signal and made it unnecessary to use a sinusoidal mechanical vibrations generator over a wide frequency range.

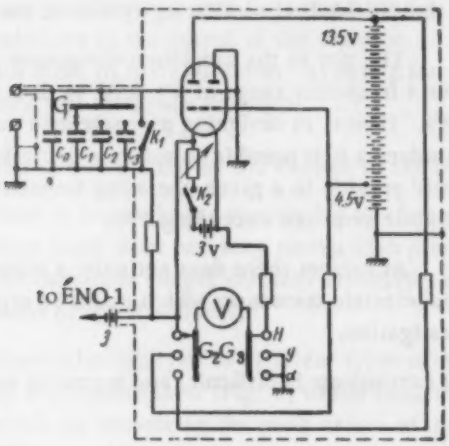

Fig. 3.

The amplitude-frequency characteristic thus obtained shows that in the range of 3-1300 cps the relative variations in the accelerometer sensitivity do not exceed ± 1%. The amplitude frequency characteristic of the ÉO-7 oscilloscope vertical amplifier rises at low frequency with a maximum at 2 cps. In the range of 3-30 cps this rise compensates the drop in the amplitude-frequency characteristic of the cathode follower input circuit. The calibration of a reference accelerometer is only possible by indirect measurements of the acceleration impressed on its transducer. Since the above accelerometer has no zero frequency characteristic, its calibration can only be made under dynamic conditions. The accuracy of dynamic calibrations is determined in the main by the accuracy of direct frequency and displacement amplitude measurements, and by the sinusoidal shape of the oscillations. For high precision of measurements it is first of all necessary to provide high frequency and amplitude stability during measurements. All these conditions can best be satisfied by exciters consisting of tuned oscillating circuits. A fixed frequency self-exciting tuning-fork generator was, therefore, used for calibrating reference accelerometers. The transducer was connected to one of the prongs of the tuning-fork and a counterpoise to the other. The displacement amplitudes in the range of 0.1-3.2 mm were evaluated by means of a measuring microscope with a relative error of 3.5-0.8%. The frequency of the tuning-fork vibrator was measured by an absolute method with an error not exceeding 0.1%.

Taking into consideration that $\delta f/\delta < 0.1\%$, the error in determining accelerations impressed on the graduated transducer was in the main determined by the accuracy in measuring the displacement amplitude $\delta A/A < 3.5\%$.

The error in calibrating the accelerometer consists of the error in determining the input signal (acceleration) and the error in determining the output signal (voltage amplitude).

The frequency of the measured voltage e may assume various values in the range of 5-1000 cps; voltage E_i produced by the calibrator has a constant frequency of 50 cps. It is possible to compare on the cathode-ray oscilloscope voltage amplitudes e and E which may have different frequencies, because the ÉO-7 oscilloscope voltage amplifier has a flat amplitude-frequency characteristic in the range under consideration. It has already been pointed out that the above equipment provides measurements of the output voltage e with an error not exceeding 1.5%.

Thus, in calibrating a reference accelerometer, i. e. in determining sensitivity $S=e/a$ we can expect to obtain values of S with an error of $\delta S=\sqrt{3.5^2+1.5^2}=3.7\%$. By increasing the number of measurements it was possible to determine the sensitivity and, hence, the calibration coefficient with an error not exceeding 0.5%. It was not con-

sidered advisable to lower the error still further, since at approximately this level the reference accelerometer may have some uneliminated systematic errors.

An analysis of the results (27 measurements) makes it possible to arrive at the following conclusions:

The quadratic mean error of the series of measurements is

$$\sigma = 0.38 \frac{mv}{g};$$

the relative error is $\sigma_0 = 1.9\%$. The quadratic mean error of the measurement result is $R = \sigma/\sqrt{n} = 0.38/\sqrt{27} = 0.07$ mv/g; and the relative error is $R_0 = 0.4\%$.

The quadratic mean error $\sigma_0 = 1.9\%$ of the series of sensitivity measurements characterizes the accuracy of this calibration method. We adopted as the calibration coefficient the mean sensitivity:

$$K = 20.4 \frac{mv}{g}.$$

The error of the reference accelerometer consists of the basic error and the dynamic error.

The basic error is determined as the error of 1.5% in comparing voltages, the error of 1.3% due to the instability of the calibrating voltage, the error of 0.7% caused by the variation in the capacitance of the cable which connects the transducer to the cathode follower, and the calibration error (maximum) of 1.2%. Thus the basic error is ± 2.5%. The dynamic error is ±1%.

The error of single measurements made by means of the reference accelerometer does not exceed ± 3.5%.

Considering that the sensitivity of the reference accelerometer transducer, and hence, the calibration coefficient of the instrument, may change with use of the instrument, it is necessary to check periodically the calibration coefficient by means of a resonance self-oscillatory system specially adapted for calibrating accelerometers at a fixed frequency.

In order to eliminate any doubt in the correctness of the reference accelerometer calibration it was decided to compare its readings with those of a specially made auxiliary accelerometer calibrated by a method based on a difference principle.

Thus, we were able to compare the results of calibrations made by two independent methods:

1) by the dynamic calibration method which consists of impressing on the accelerometer transducer sinusoidal vibrations, and calculating from direct measurements of the frequency and displacement amplitudes the amplitude of the transducer's acceleration, and

2) by the static calibration method, which consists in applying to the inertial mass of the transducer's accelerometer sensitive element with a zero frequency characteristic a force which bears a certain relation to the weight of that mass. In order to be able to calibrate the auxiliary accelerometer statically and then compare its readings with those of the reference accelerometer in a dynamic condition, it was necessary to provide the auxiliary accelerometer with a constant sensitivity in the frequency range of 0-250 cps. This requirement is met by an accelerometer with a piezoelectric quartz transducer and a cathode follower with an electrometric tube. The block schematic of the auxiliary accelerometer is similar to that of the reference one. It consists of a transducer with a piezoelectric quartz sensitive elements, a cathode follower with an electrometric tube 2Ē2P and a cathode-ray oscilloscope ĖNO-1.

Quartz plates 20 mm in diameter and 4.5 mm thick were used in the auxiliary accelerometer transducer. Polystyrene insulation was used for the output leads. The upper part of the transducer has a device for fixing coaxially with it the reference accelerometer transducer when they are being compared.

The transducer's inertial mass is 750 g, it is made of steel, the transducer's total weight is 2160 g, its sensitive element's natural frequency is $\omega_0 = 6200$ cps.

The schematic of the auxiliary accelerometer's cathode follower is given in Fig. 3.

The insulation resistance of the transducer line and that of the tube grid is of the order of 10^{13} ohm. Capaci-

tors C_0,C_1,C_2 and C_3 have polystyrene insulation and serve to vary the transducer's sensitivity. The capacitors, the tube and the power supplies are screened. The grid of the tube when grounded has a potential of 3 v with respect to the cathode. In the operating condition of the accelerometer the control grid potential with respect to the cathode varies between 2-4 v.

The output voltage is taken off the electrometric tube cathode. In order to balance out the direct component of the output voltage a 3 v battery is connected in series with the output. In order to avoid the cable effect, the cable from the transducer is connected to the electrometer tube grid through a screened spiral spring. The output voltages were compared by means of oscilloscope ÉNO-1, which has a vertical amplifier with a zero frequency characteristic, and incorporates a voltage calibrator. The auxiliary accelerometer had four ranges of 0-4 g (sensitivity of 15.5 mv/g), 0-9 g (6.66 mv/g), 0-20 g (3.13 mv/g) and 0-30 g (2.10 mv/g). Since the ÉNO-1 oscilloscope vertical amplifier has a smaller range of gain control than oscilloscope ÉO-7, the relative error in comparing output voltages, which determined in the main the error of the auxiliary accelerometer readings, was in the range of ± 4%.

The readings of the reference and auxiliary accelerometers were compared on mechanical vibration racks which did not provide sinusoidal vibrations.

For these comparisons the auxiliary accelerometer transducer was fixed by means of bolts to the vibration racke platform and the reference accelerometer transducer screwed on to the upper surface of the auxiliary transducer in a position where their axes coincided.

An analysis of the results obtained in comparing the reference and the auxiliary accelerometers leads to the following conclusions.

The difference in the readings of the two instruments varies in sign and magnitude in the range of ± 4% of the measured value. Of the 27 differences 13 have a positive and 14 a negative sign $\Sigma\Delta a_i = 24.4$, $\Sigma-\Delta a_i = -23.9$. No systematic difference between the readings of the reference and auxiliary accelerometers was observed. This circumstance confirms the validity of either of the calibration methods.

The quadratic mean value of the difference in readings $R = \sqrt{\frac{\Sigma(\Delta a_i)^2}{n\,(n-1)}} = 0.4\%$; hence in calibrating commercial accelerometers by comparing them with the reference accelerometer it is possible to reduce by repeated measurements the effect of radom errors on the measurement results, that is,on the calibration coefficient.

The production by the appropriate organizations of reference accelerometers constructed in a unit form and of the equipment required for their calibration consisting of resonance self-oscillating vibrators similar to those described in this article, will greatly simplify the calibration of accelerometer equipment and raise the accuracy of its readings.

EQUIPMENT FOR DYNAMIC TESTING OF PRESSURE TRANSDUCERS

A. P. Pleshko and V. V. Perfil'ev

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pp. 17-19, July, 1961

Pressure transducers with a low inertia are widely used in testing motors and other machines. The readings of these transducers are evaluated on the basis of their static calibration.

The transducers normally register both the direct and the alternating component of pressure. Hence the total error which includes that of the transducer itself, the converter and the oscilloscope amounts to 5-7% of P_{nom}.

The dynamic errors of the devices under consideration can be evaluated reliably by testing them on an equipment which can provide pressures varying according to a given law.

Such an equipment suitable for dynamic testing of transducers intended for recording only the alternating

component of pressure is described below (Figs. 1 and 2).

The pressure amplitude ΔP depends on mass m of the oscillating mercury, on the vibrational acceleration amplitude a and area S which is covered by the mercury

$$\Delta P = \frac{ma}{S} = \frac{PNg}{gS} = \frac{PN}{S}, \quad (1)$$

where P is the weight of the oscillating mass, kg; S is the area affected by the alternating force, cm^2; $N = A \cdot 4\pi^2 f^2$ is the vibrational acceleration amplitude in units of g.

If the area of the vibrating mercury column is equal to that covered by the mercury we have

$$\Delta P = \gamma \cdot l \cdot N, \quad (2)$$

where γ is the specific gravity of mercury, kg-wt/cm^3; l is the height of the mercury column, cm.

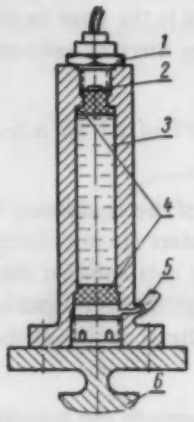

Fig. 1. Schematic of the equipment. 1) pressure transducer; 2) transducer diaphragm; 3) vibrating mass (mercury); 4) rubber packing; 5) static air pressure supply; 6) vibrator table.

Reliable operation and sufficiently high accuracy of the above equipment is provided by the following conditions.

1. The amplitude of the vibrator oscillations must be several times as great as the sagging of the transducer diaphragm, since only under this condition will the displacement of the mercury with respect to the casing be sufficiently small.

2. The area of the transducer diaphragm must be only a small fraction of the vibrating mercury column area.

3. The height of the mercury column must be

$$l < \frac{1}{4}\lambda < \frac{w}{4f}, \quad (3)$$

where ω is the speed of sound in mercury.

4. The frequency of the measured oscillations must be at least 5 times greater than the instrument's resonant frequency.

5. The vibrational accelerations must not exceed 300-350 g which are normally used in testing the shock-resistant and vibration-proof properties of transducers, i. e. the set vibrational accelerations must not impede the normal operation of the transducer.

6. The device must be suitable for testing with a static pressure

$$P \approx 50\text{—}70\% \text{ of } P_{nom}.$$

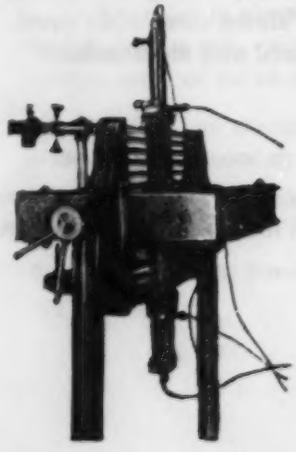

Fig. 2. General appearance of the equipment.

In order to obtain maximum pressure pulsation amplitudes the above equipment uses mercury as the vibrating body, and a frame-vibrator as a test rack which provides vibrational accelerations up to 125 g. The test rack frequency of vibrations is f = 250 cps. The height of the oscillating mercury column is l = 60 mm, which is 1/25 of the oscillatory quarterwavelength (λ), since for f = 250 cps and $\omega = 1.45 \cdot 10^3$ m/sec we have $\frac{1}{4}\lambda = 1.45$ m. The internal diameter of the device (diameter of the vibrating mercury column) d = 20 mm. The value of the double vibration amplitude of the frame-type test rack is 2A=1.0 mm, which is several hundred times larger than the displacement of the mercury column due to the sagging of the transducer's sensitive element (taking into account the ratio K=4 between cross sectional area S_p of the vibrating mercury column and the area S_d of the

transducer's sensing element). The natural frequency of the device with the transducer screwed into position was determined experimentally and amounted to f ≈ 2250 cps, which is about 9 times the frequency of the set vibrations. The shape of the pressure pulsation curve thus produced was almost completely sinusoidal.

It was found by means of experimental investigations that the equipment provides a pressure pulsation amplitude of P ≈ ± 10 atm at a frequency of f = 250 cps.

The error in evaluating the established pressure pulsation amplitude is determined from formula:

$$\delta\Delta P = \delta P + \delta(2A) + 2\delta f + 2\delta d, \qquad (4)$$

where δ P is the error in determining the weight of the vibrating mercury column; δ (2A) is the error in determining the double vibration amplitude: δ f is the error in determining the frequency of vibrations; δ d is the error in determining the diameter of the vibrating mercury column (the internal diameter of the device).

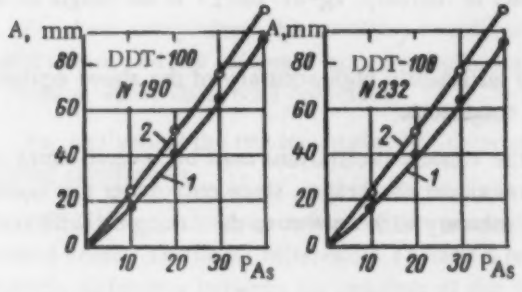

Fig. 3. Calibration curves of two strain gauge pressure transducers. 1) Calibrated on the device; 2) calibrated on the press.

This error does not exceed 0.9% of ΔP for a double vibration amplitude of 1.0 mm.

During the experimental use of the equipment it was established that the vibrator parameters did not change when its frame was rotated about its axis and that the frequency of the pressure pulsations varied over a considerable range (~ 10-15 cps) depending on the accuracy of the vibrator tuning.

Accuracy in transmitting pressure to the transducer's sensing elements was determined by testing strain gauge transducers and comparing the results with their calibration characteristics obtained in the normal way on a press. The pressure transducers were used in conjunction with a 6-channel strain gauge equipment 6-TU-35-ShA, and a type 4 loop of a Siemens oscilloscope was used for recording purposes.

It will be seen from the calibration curves shown in Fig. 3 that at a pressure of 10 atm they exhibit a small difference, probably due to the instability of normal tensions on the rubber-diaphragm boundary. At higher pressures, however, this tendency disappears and the slope angles of the calibrating curves become almost completely equal. In order to eliminate this error in the strain gauge readings a calibration curve was obtained with the transducer connected directly to the vibration equipment.

CONCLUSIONS

The above equipment has made it possible not only to determine the errors of systems used for recording pulsating pressures, but also to test transducers under conditions when they have been subjected simultaneously to pressure pulsations and vibrational accelerations.

ELECTRODYNAMIC DEFLECTOMETER

G. S. Rodov

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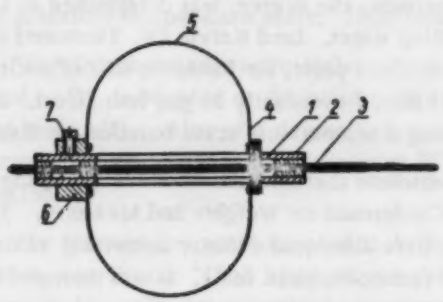
For the purpose of measuring sagging in dynamic testing of building constructions, several instruments have been produced on the basis of strain-gauge wire transducers. The deflectometer developed by S. A. Grach [1] is of interest. It is, however, not free from certain defects. When the absolute values of sagging become large the readings of this instrument become unstable due to its one-sided elastic strain gauge bow asymmetrical with respect to the rod, thus producing a skewing and wedging of the rod, especially with a slanting applied effort.



The West Siberian Branch of the Building and Architectural Academy, USSR has developed under the guidance of the author of this article an improved electrodynamic deflectometer with a double strain gauge bow.* Basic changes introduced into the design of this instrument as compared with the design of S. A. Grach eliminate the above mentioned defects and considerably raise the stability of its readings. The instrument (see Figure) consists of a frame 1, in which rod 3 moves in guide bushes 2. Spring steel bow strips which are fixed to the rod form semicircular beams 4 with strain gauges 5. The other end of the steel bows (beams) is fixed to sleeve 6 which can move along frame 1 for setting the sensitivity of the instrument. The sleeve is secured by means of a clamping screw 7. The existence of two semicircular symmetrical beams prevents skewing in the displacement of the rod. The strain gauge transducers which are glued to the beams transform into electrical quantities the relative deformations of the beams due to the displacements of the rod with respect to the frame which arise in measuring sagging.

The electrodynamic deflectometer is placed at the point where the sagging has to be measured and fixed by means of a lug in its tubular frame to a solid base, and the strain gauge wire transducers are connected to a measuring bridge or an oscilloscope.

The error in measuring the sagging depends on the error of evaluating relative deformations by means of the bridge circuit, and varies between 0.001 and 0.1 mm per scale calibration of the measuring instrument. The calibration curve of the electrodynamic deflectometer approaches a straight line.

In 1960 the electrodynamic deflectometer was exhibited at the "Building" section of the Exhibition of the USSR National Economy Achievements. The Exhibition commission recommended mass production of the instrument.

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1. S. A. Grach, Izmeritel'naya Tekhnika 5 (1958).

* V. A. Voznesenskii and N. N. Federov participated in the development work.

THERMOTECHNICAL MEASUREMENTS

NEW REGULATIONS ON THE INTERNATIONAL PRACTICAL TEMPERATURE SCALE

B. N. Oleinik

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This article deals with problems related to the adoption in October, 1960, by the 11th General Conference on Weights and Measures of the document entitled "1948 Regulation on the International Practical Temperature Scale. 1960 Edition."

The absolute thermodynamic temperature scale proposed by Kelvin in 1854 is the original scale on which all modern temperature scales are based. In this scale the unit of measurement, the degree, was determined as 1/100 of the interval between the temperature of melting ice and that of boiling water. Lord Kelvin (W. Thomson) considered that the determination of the temperature scale by means of one fixed point, for instance, that of melting ice, to which a definite numerical value is allocated, is preferable and should eventually be put into effect. In 1873 D. I. Mendeleev expressed a similar idea on the need of introducing a temperature scale based on one fixed point.

Experimental difficulties connected with the introduction of an absolute thermodynamic scale led to the adoption of an international temperature scale in 1927 by the 7th General Conference on Weights and Measures. This scale is based on fixed points of phase equilibrium temperatures which were allocated definite numerical values. The 1927 scale was designed to coincide as nearly as possible with the thermodynamic scale. It was arranged in a manner to reproduce temperatures with an error smaller than that of the thermodynamic scale.

The first modification of the 1927 International Temperature Scale was made by the 9th General Conference on Weights and Measures which approved the "1948 Regulation on the International Temperature Scale." The 1948 Regulation noted the basic differences in the methods used for reproducing the practical temperature scale as compared with the methods established by the 1927 Regulation.

"The methods by which the new scale should be reproduced remain essentially the same as those used for the 1927 scale. Only two changes in specifying the scale produce an appreciable difference in the numerical value of the measured temperatures, namely: 1) the amendment in the numerical value of the silver point (960.8°C in place of 960.5°C) changes all the temperatures measured by thermocouples; 2) the establishment of a new value for the radiation constant C_2 changes all the temperatures above the solidification point of gold, whereas the use of Planck's instead of Wien's formula affects mainly only very high temperatures. Planck's formula makes the new scale more comparable to the thermodynamic scale, since it eliminates the upper limit of the scale which in the 1927 scale was fixed by Wien's law.

"Other important changes of the 1927 scale which produce only small variations in the numerical values of temperatures or even do not alter them at all, but provide a clearer definition and better reproducibility of the temperature scale are the following:

" a) fixing the scale lower limit at the oxygen point (and not at point −190°C used in the 1927 scale).

" b) changes in the scale subdivisions, namely: specifying the antimony point (about 630°C) instead of the point at 660°C for the beginning of the third scale section (from the antimony to the gold point);

" c) more stringent requirements for the purity of platinum used for making reference resistance thermometers and reference thermocouples, as well as narrower tolerances for the variations in the thermal-emf of reference thermocouples at the gold point.

"the scale sections determined by resistance thermometers remain essentially the same as those in the 1927 scale. In the range of 630 to 1063°C the numerical values of the 1948 temperature scale are higher than those of the 1927 scale, the maximum difference at 800°C is approximately 0.4°.

"The 1927 international scale temperatures were denoted as °C or °C (Int.). Since the notation '°C' is preserved in the new regulation it will henceforth be necessary to append to the above notations the year of the scale, i. e.

in cases when any doubt can arise the notation must be specified in the following manner: °C (Int. 1927) or °C (Int. 1948)."

In the 1948 regulation it is again stressed that Kelvin's scale is "recognized as the basic thermodynamic scale in which any temperature measurement can be finally expressed. In this scale the temperature interval between the melding point of ice T_0 and the boiling point of water T_{100} is equal to 100 degrees. By the present resolution the 9th General Conference on Weights and Measures adopts side by side with Kelvin's scale a centigrade thermodynamic scale, whose temperature is equal to $T-T_0$. In these scales any temperature interval will have the same numerical value."

In 1954 the 10th General Conference on Weights and Measures adopted a new definition of the thermodynamic temperature scale.

This scale with its beginning at the absolute zero has only one basic fixed point, the triple point of water, which determines the numerical value of a degree.

The new definition of the absolute thermodynamic scale is reflected in the "1948 Regulation on the International practical temperature scale. 1960 Edition, "adopted by the 11th General Conference on Weights and Measures:

"Kelvin's thermodynamic scale, whose temperatures are denoted by symbol T and whose numerical values are denoted by °K, is adopted as the basic scale; there should be the possibility to refer any temperature measurement to that scale. The value of a degree Kelvin is determined by the decision which established the thermodynamic temperature of the triple point of water at 273.16°K (10th General Conference on Weights and Measures, 1954, Resolution 3)."

The temperature scale adopted in 1960 under a new name by the International Committee on Weights and Measures, namely the "1948 International practical temperature scale. 1960 Edition" does not alter the 1948 temperature scale, since the numerical temperature values determining the scale remain the same as in 1948. The 1948 Regulation, 1960 Edition, is a document which repeats the "1948 Regulation on the International temperature scale" in a different wording.

In the 1948 Regulation, 1960 Edition, the following definition of the International practical temperature scale is established:

"Temperatures in the 1948 International Practical scale are expressed in degrees Celsius, and denoted by °C or °C (Int. 1948), and are represented by symbols $\underline{t}$ or t_{int}.

"The International practical temperature scale is based on six reproducible temperatures (primary fixed points), which have been allocated numerical values, and on formulas giving the relation between the temperature and the readings of instruments calibrated by means of these six primary fixed points. These fixed points are determined by states of equilibrium obtained according to specification; with the exception of the triple point of water these states of equilibrium are taken with a pressure of 101,325 newtons per square meter (normal atmosphere).

"The primary fixed points of the scale and precise numerical values which have been allocated to them are given in Table 1.

The recommendation to use the solidification point of zinc instead of the boiling point of sulfur is based on investigations carried out in the metrological organizations of the USSR, Canada, USA and Britain. In the USSR the above research was carried out by the VNIIM (All-Union Scientific Research Institute of Metrology).

According to the interpolation methods the temperature scale is divided into four ranges:

a) from 0 to 630.5°C (solidification point of antimony);

b) from the boiling point of oxygen to 0°C;

c) from 630.5°C to the solidification point of gold;

d) above the solidification point of gold.

The scale is reproduced in the first two ranges by means of platinum resistance thermometers, in the third range by means of platinum-platinorhodium thermocouples, and in the fourth by means of optical pyrometers.

The instruments, methods an techniques recommended by the Regulation reflect the level of the knowledge attained at the present stage of development. These recommendations improve uniformity and reproducibility in applying the International Practical Temperature Scale.

TABLE 1. Primary Fixed Points

Precision values allocated to them. The pressure is equal to 1 normal atmosphere, with the exception of the triple point of water

	Temperature °C(Int. 1948)
Temperature of equilibrium between liquid oxygen and its vapor (boiling point of oxygen)	- 182.97
Temperature of equilibrium between ice, water and water vapor (triple point)	+ 0.01
Temperature of equilibrium between water and its vapor (boiling point of water)	100
Temperature of equilibrium between liquid sulfur and its vapor (boiling point of sulfur)	444.6*
Temperature of equilibrium between solid and liquid silver (solidification point of silver)	960.8
Temperature of equilibrium between solid and liquid gold (solidification point of gold)	1063 "

* Instead of the boiling point of sulfur it is recommended to use the point of temperature equilibrium between solid and liquid zinc (solidification point of zinc) which is allocated the value of 419.505° C (Int. 1948). This point is easier to reproduce than that of sulfur, and its value has been chosen so that it provides the same result in determining temperature by the International Practical Scale.

The Regulation recommends the use in reproducing the scale not only of primary but also of secondary fixed points:

"In addition to the primary fixed points which determine the scale and are given in Table 1, it is possible to use other reference points. Some of these points with their temperatures given in the 1948 International Practical Temperature Scale are listed in Table 2. With the exception of the triple points all the remaining points consist of temperature systems in a state of equilibrium at a pressure of 1 normal atmosphere. Formulas for the relation between temperature and pressure are given for the range of p=680 to p=760 mm Hg.

The regulation also examines the relation between the International Temperature Scale and the Thermodynamic Scale. Certain experimental data on the differences between these scales are given for the sake of information, in order to show the present level of our knowledge in this respect.

Table 3 gives the basic recommendations on the notation of temperature and temperature units.

Details of the temperature scale history and their peculiarities will be found in works [1,2,3].

The basic requirements of the '1948 Regulation on the International Practical Scale. 1960 Edition' are taken into account in the new GOST (All-Union State Standard) now being compiled on 'Thermal Units' in order to replace GOST 8550-57 on 'Thermal Units' and OST VKS (All-Union Standar of the All-Union Standardization Committee(6954 on the 'International Temperature Scale.'

TABLE 2. Secondary Fixed Points at a Pressure of 1 Normal Atmosphere (with the Exception of Triple Points)

	Temperature °C (Int. 1948)
Temperature of equilibrium between solid carbon dioxide and its vapor	-78.5
$t_p = \left[-78.5 + 12.12\left(\frac{p}{p_0} - 1\right) - 6.4\left(\frac{p}{p_0} - 1\right)^2\right]$ °C	
Temperature of equilibrium between solid and liquid mercury	-38.87
Temperature of equilibrium between ice and water saturated with air	0.000
Temperature of the triple point of phenoxybenzene (diphenyl oxide)	26.88
Transition temperature of sodium sulfate decahydrate	32.38
Temperature of the triple point of benzoic acid	122.36
Temperature of equilibrium between solid and liquid indium	156.61
Temperature of equilibrium between liquid naphthalene and its vapor	218.0
$t_p = \left[218.0 + 44.4\left(\frac{p}{p_0} - 1\right) - 19\left(\frac{p}{p_0} - 1\right)^2\right]$ °C	
Temperature of equilibrium between solid and liquid tin	231.91
Temperature of equilibrium between liquid benzophenone and its vapor	305.9
$t_p = \left[305.9 + 48.8\left(\frac{p}{p_0} - 1\right) - 21\left(\frac{p}{p_0} - 1\right)^2\right]$ °C	
Temperature of equilibrium between solid and liquid cadmium	321.08
Temperature of equilibrium between solid and liquid lead	327.3
Temperature of equilibrium between mercury and its vapor	356.58
$t_p = \left[356.58 + 55.552\left(\frac{p}{p_0} - 1\right) - 23.03\left(\frac{p}{p_0} - 1\right)^2 + 14.0\left(\frac{p}{p_0} - 1\right)^3\right]$ °C	
Temperature of equilibrium between solid and liquid aluminum	660.1
Temperature of equilibrium between solid and liquid copper (in a reducing atmosphere)	1083
Temperature of equilibrium between solid and liquid nickel	1453
Temperature of equilibrium between solid an liquid cobalt	1492
Temperature of equilibrium between solid and liquid palladium	1552
Temperature of equilibrium between solid and liquid platinum	1769
Temperature of equilibrium between solid and liquid rhodium	1960
Temperature of equilibrium between solid and liquid iridium	2443
Fusing temperature of tungsten	3380.*

TABLE 3. International Practical Temperature Scale

International practical temperature	International practical temperature (Absolute)
t_{int}	$T_{int} = t_{int} + T_0$
Unit notation: °C (Int. 1948)	Unit notation: °K (Int. 1948)
Degree Celsius international practical 1948	Degree Kelvin, international practical 1948

TABLE 3 (continued)

Thermodynamic Temperature Scale

Thermodynamic temperature (in degrees Celsius)	Thermodynamic temperature (absolute)
$t = T - T_0$	T
Unit notation: °C (therm)	°K
Degree Celsius thermodynamic	Degree Kelvin

($T_0 = 273.15°$)

Note:

1) In the notation of the International Practical Temperature subscript "int" need not be appended to t if this does not cause any ambiguity.

2) The arrows indicate the direction of transition from a temperature determined originally to the temperature obtained by changing the beginning of reckoning.

The full text of the Regulation will be published in the current collection of the VNIIM Works entitled "Research in the sphere of thermal measurements."

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ELECTRICAL MEASUREMENTS

TEMPERATURE COMPENSATION OF TRANSISTOR STABILIZERS

S. D. Dodik

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Transistor stabilizers are being increasingly used in radioelectronics, automation and measurement technology as powerful sources of reference voltage which is comparable in its stability with that of standard cells [1, 2, 3]. It has been shown in [2] that by means of a simple circuit it is possible to obtain a total instability of the order of 0.01% and even higher with both supply voltage and load current variations. In stabilizers which use as their voltage reference source silicon stabilitrons (KS) [4,5,6,7] the main cause of temperature instability consists in the drift of the reference voltage and that of the transistor base voltage of the dc amplifier (DCA) used in the comparison circuit. In the case of high-voltage feedback potential dividers an important part is played by the zero collector current i_{co} of the DCA (germanium) transistor, whose effect, however, can be neglected if a low-resistance divider is used. The potential divider resistors may introduce a large error if they are not made of manganin. The part played by current i_{co} and the variations in the voltage of the control transistor base is considerably smaller, since these effects are reduced by a large feedback.

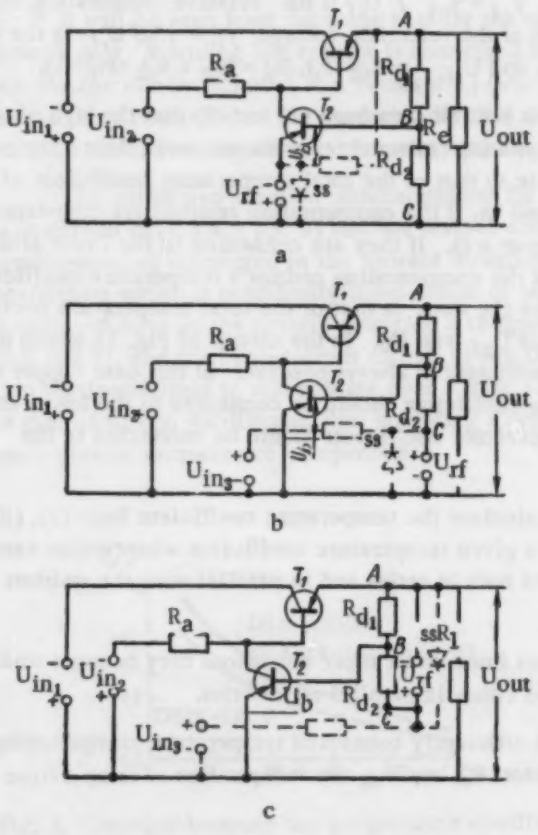

Fig. 1. Basic transistor stabilizer circuits.

Let us now examine three basic transistor stabilizer circuits with a single-stage DCA (Fig. 1 a,b,c) in order to develop methods for their temperature compensation. These circuits differ by the method of connecting the source of the reference voltage U_{ref}. In Fig. 1a, described in [2] the output voltage U_{out} can only be larger than U_{ref}, in the circuit of Fig. 1b it can be both smaller and larger, and in the circuit of Fig. 1c it can only be smaller than U_{ref}.

In Fig. 1, T_1 is the controlling transistor (it can be compound) and T_2 is the DCA transistor.

U_{in1}, U_{in2}, U_{ref} and U_{out} are the controlling transistor and DCA supply voltages, and the reference and output voltages;

R_a, R_1, R_{d1} and R_{d2} are the DCA and stabilizer load resistors, and the upper and lower arms of the feedback potential divider.

The silicon stabilitron shown by a dotted line on the drawings of Figs 1b and c requires an additional voltage source U_{in3}.

The absolute temperature coefficient (TC) of the silicon stabilitrons D808-D813 [6,7] is γ'_{ref} (in mv/°C), it remains positive and practically constant in the temperature range of -40 to +50°C. Stabilitrons with an absolute temperature coefficient approaching zero or even becoming negative are theoretically possible [4,5] and have been obtained in practice. The absolute voltage TC between the DCA transistor base and emitter denoted by γ'_b is negative and equal to some 2 mv/°C for an emitter current of 3-5 ma. The stabilizer TC can be, depending on the circuit components, either positive or negative.

The transistor stabilizers are best compensated for temperature in the DCA base circuit by including in the

$$R_{d_1}=R'_{d_1}+R''_{d_1}\quad R_{d_2}=R'_{d_2}+R''_{d_2}\quad R_{d_1}=\frac{R'_{d_1}R''_{d_1}}{R'_{d_1}+R''_{d_1}}\quad R_{d_2}=\frac{R'_{d_2}R''_{d_2}}{R'_{d_2}+R''_{d_2}}$$

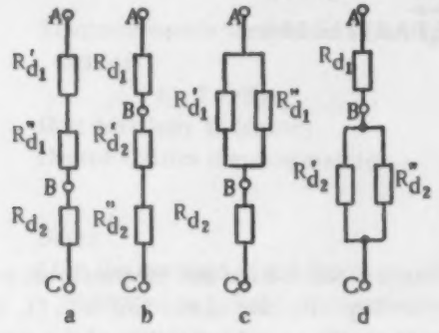

Fig. 2. Circuits of thermal compensation by means of temperature sensitive resistors.

$$R_{d_1}=\frac{R'_{d_1}R''_{d_1}}{R'_{d_1}+R''_{d_1}}\quad R_{d_2}=\frac{R'_{d_2}R''_{d_2}}{R'_{d_2}+R''_{d_2}}\quad R_{d_2}=\frac{R'_{d_2}R''_{d_2}}{R'_{d_2}+R''_{d_2}}\quad R_{d_1}=\frac{R'_{d_1}R''_{d_1}}{R'_{d_1}+R''_{d_1}}$$

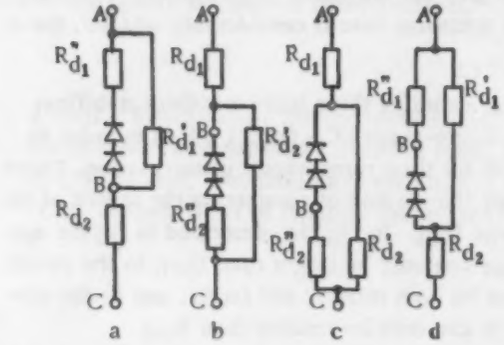

Fig. 3. Circuits for temperature compensation by means of p-n-junction semiconductors.

potential divider temperature-sensitive resistors [8] or semiconductors with a forward p-n junction [2,3,9].

Depending on the sign of the stabilizer resistance temperature drift and that of the temperature coefficient of the resistor, the latter is connected either into the upper or the lower arm of the feedback potential divider circuit. It can be shown that in the circuit of Fig. 1a, in which resistor R_{d1} and R_{d2} depend on temperature it is necessary in order to compensate the temperature coefficient to have a compensating resistor γ which meets the following requirements.

If the compensating resistor is connected to the upper arm of the divider:

$$\gamma=-\frac{\gamma_{rf}+\gamma_b}{1-n}. \tag{1}$$

If it is connected to the lower arm:

$$\gamma=\frac{\gamma_{rf}+\gamma_b}{1-n}, \tag{2}$$

where $\gamma_{rf}=\gamma_{rf}/U_{rf}$ is the relative temperature coefficient of the reference voltage; $\gamma_b=\gamma'_b/U_{rf}$ is the ratio of γ'_b and U_{rf}; $n=R_{d2}/R_d$, $R_d=R_{d1}+R_{d2}$ (Fig. 1).

It will be seen from (1) and (2) that the sign of the compensating resistors' temperature coefficient must be opposite to that of the total temperature coefficient of U_{rf} and U_b if the compensating resistors are connected to the upper arm. If they are connected to the lower arm the sign of the compensating resistor's temperature coefficient must be the same as that of the total temperature coefficient of U_{rf} and U_b. In the circuit of Fig. 1a which uses silicon stabilitrons D808-D810 the total voltage temperature coefficient is always positive. In this case copper or nickel compensating resistors which have a positive temperature coefficient should be connected to the lower arm, but standard thermal compensation resistors with a negative temperature coefficient should be connected to the upper arm.

In practice it is usually impossible and not advisable to calculate the temperature coefficient from (1), (2) for all the resistors R_{d1} and R_d. In practice there are resistors with a given temperature coefficient whose value can be varied. For this purpose compensating resistors can be connected both in series and in parallel with the resistors unaffected by temperature.

Series connection is always preferable for metallic resistors since under those conditions they become small with a small time constant. Semiconductor resistors can be used either in parallel or in series.

Figure 2a, b, c, and d show potential divider circuits with differently connected temperature-compensating resistors R''_{d1} and R''_{d2}. Resistors R'_{d1} and R'_{d2} which comprise resistors R_{d1} and R_{d2} are independent of temperature variations.

An analysis provides the following conditions of temperature compensation for the circuits shown in Fig. 1a:

$$\text{Fig. 2 a: } R''_{d1}=-R_d\frac{\gamma_{rf}+\gamma_b}{\gamma} \tag{3}$$

$$\text{Fig. 2 b: } R''_{d2}=R_d\frac{n}{1-n}\cdot\frac{\gamma_{rf}+\gamma_b}{\gamma} \tag{4}$$

Fig. 2 c: $R''_{d1} = -R_d (1-n)^2 \frac{\gamma}{\gamma_{rf} + \gamma_b}$ (5)

Fig. 2 d: $R''_{d2} = R_d (1-n)n \frac{\gamma}{\gamma_{rf} + \gamma_b}$ (6)

In the case of stabilizers shown in Figs. 1b and 1c, the formulas for R''_{d1} and R''_{d2} have the form shown in the Table (we give them without derivations).

The total divider resistance R_d is selected on the basis of minimum instability [2] with a changing load and supply voltage, as well as for a minimum effect of current i_{rf}, the value of n is selected so as to obtain the required output voltage for a given reference voltage, γ'_b is normally about 2 mv/°C, and hence, with the knowledge of U_{rf} it is easy to find γ_b. If silicon stabilitrons D808 - D813 are used, the value of γ_{rf} is determined from the reference voltage and the curve shown in [6] or [7], and the value of γ from reference books.

Thermistors KMT and MMT have a negative temperature coefficient of the order of 2.4 - 3.4%/°C (MMT) and 4.5 - 6%/°C (KMT), which is inversely proportional to the square of the absolute temperature. For MMT thermistors the temperature coefficient drops at the rate of $\frac{0.02\%/°C}{°C}$, and for thermistors KMT at the rate of $\frac{0.035\%/°C}{°C}$. Therefore, when it is necessary to compensate for temperature over a wide range, or to compensate with precision, the use of thermistors is not advisable. In such cases it is preferable to use metallic copper or nickel resistors with a temperature coefficient of 0.39%/°C and 0.5%/°C respectively. Since the resistivity of nickel is about 6 times as great as that of copper and its temperature coefficient is about 1.3 times greater than that of copper, nickel resistors used for temperature compensation will be 1/8 the size of copper resistors with the remaining conditions being equal. Moreover, the thermal inertia of such a resistor is reduced and its temperature compensation improved.

It will be seen from the table that for the circuit in Fig. 1a the value of the compensating resistor does not depend on n, providing the resistor is connected to the upper arm of the potential divider. This means that the output voltage can be adjusted at a constant R_d (which can be done by connecting the base of T_1 to the midpoint of the variable resistance of the potential divider) without interfering with the temperature compensation of the circuit. In the remaining circuits temperature compensation is set only for a given output voltage.

The use of p-n-junction semiconductors for compensating the temperature coefficient of silicon stabilitrons is described in [2,3,4,7 and 9] and has several advantages over temperature-sensitive resistors. Thus, the p-n-junction semiconductors connected in the forward direction have at currents greater than 5 ma an absolute temperature coefficient which is practically independent of temperature over a wide range (from -40 to +50°C). Since the silicon stabilitrons have in the reverse direction a temperature coefficient which is also little dependent on temperature [4,5 and 7], and the base-emitter circuit of the DCA also represents a p-n junction, it is possible by using p-n-junction semiconductors to compensate over a wide temperature range. The inertia of p-n junctions is almost the same as that of silicon stabilitrons, thus providing a synchronized variation of the voltage with temperature, and hence a more precise temperature compensation.

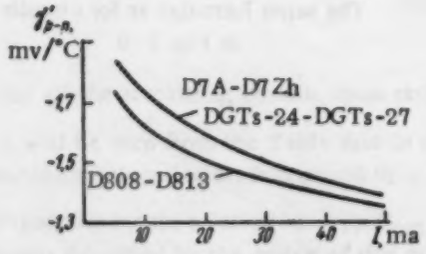

Fig. 4. Relation between the temperature coefficient of p-n-junction semiconductors and current.

It has been shown in [2,3] that compensating p-n-junction semiconductors should be included in a potential divider circuit instead of being connected in series with the basic silicon stabilitrons. The p-n-junction semiconductors connected in different arms of the potential divider can compensate both positive an negative temperature coefficients of the stabilizer. Figures 3a,b,c,d show the manner of connecting compensating p-n junction semiconductors in the potential divider circuit. Since the temperature coefficient voltages of the p-n-junction semiconductors connected in the forward direction are always negative, circuits of Figs. 3a and c and Figs. 3b and d are used respectively for the total positive and negative temperature coefficients of the uncompensated stabilizer.

The Table provides formulas, without deriving them, for determining the number of compensating p-n junction-semiconductors required for the respective circuits. These formulas do not account for the effect of the dynamic

Conditions for Temperature Compensation of Transistor Stabilizers

Stabilizer circuit	Temp.-compensating circuit	Conditions of temperature compensation when compensating resistors are used	Temp.-compensating circuit	Conditions of temperature compensation when p-n-junction semiconductors are used
Circuit of Fig. 1 a	Fig. 2 a	$R''_{d1} = -R_d \frac{\gamma_{rt} + \gamma_b}{\gamma}$	Fig. 3 a	$\gamma'_{p-n}N = -\frac{\gamma'_{rt} + \gamma'_b}{n} \Big/ \left[1 - \frac{R_d(1-n)}{R'_{d1}}\right]$
	Fig. 2 b	$R''_{d2} = R_d \frac{n}{1-n} \frac{\gamma_{rt} + \gamma_b}{\gamma}$	Fig. 3 b	$\gamma'_{p-n}N = \frac{\gamma'_{rt} + \gamma'_b}{1-n} \Big/ \left(1 - \frac{R_d n}{R'_{d2}}\right)$
	Fig. 2 c	$R''_{d1} = -R_d(1-n)^2 \frac{\gamma}{\gamma_{rt} + \gamma_b}$	Fig. 3 c	$\gamma'_{p-n}N = -\frac{\gamma'_{rt} + \gamma'_b}{n} \Big/ \left[1 + (1-n)\frac{R_d}{R'_{d2}}\right]$
	Fig. 2 d	$R''_{d2} = R_d(1-n)n \frac{\gamma}{\gamma_{rt} + \gamma_b}$	Fig. 3 d	$\gamma'_{p-n}N = \frac{\gamma'_{rt} + \gamma'_b}{1-n} \Big/ \left(1 + \frac{R_d n}{R'_{d1}}\right)$
Circuit of Fig. 1 b	Fig. 2 a	$R''_{d1} = -R_d \frac{\gamma_{rt}(1-n) + \gamma_b}{\gamma}$	Fig. 3 a	$\gamma'_{p-n}N = -\frac{\gamma'_{rt}(1-n) + \gamma'_b}{n} \Big/ \left[1 - \frac{R_d(1-n)}{R'_{d1}}\right]$
	Fig. 2 b	$R''_{d2} = R_d \frac{n}{1-n} \frac{\gamma_{rt}(1-n) + \gamma_b}{\gamma}$	Fig. 3 b	$\gamma'_{p-n}N = \frac{\gamma'_{rt}(1-n) + \gamma'_b}{(1-n)} \Big/ \left(1 - \frac{R_d n}{R'_{d2}}\right)$
	Fig. 2 c	$R''_{d1} = -R_d(1-n)^2 \frac{\gamma}{\gamma_{rt}(1-n) + \gamma_b}$	Fig. 3 c	$\gamma'_{p-n}N = -\frac{\gamma'_{rt}(1-n) + \gamma'_b}{n} \Big/ \left[1 + (1-n)\frac{R_d}{R'_{d2}}\right]$
	Fig. 2 d	$R''_{d2} = R_d(1-n)n \frac{\gamma}{\gamma_{rt}(1-n) + \gamma_b}$	Fig. 3 d	$\gamma'_{p-n}N = \frac{\gamma'_{rt}(1-n) + \gamma'_b}{(1-n)} \Big/ \left(1 + \frac{R_d n}{R'_{d1}}\right)$
Circuit of Fig. 1 c	Fig. 2 a	$R''_{d1} = -\frac{R_d}{n} \frac{\gamma_{rt}(1-n) + \gamma_b}{\gamma}$	The same formulas as for circuits in Fig. 1 b	
	Fig. 2 b	$R''_{d2} = \frac{R_d}{1-n} \frac{\gamma_{rt}(1-n) + \gamma_b}{\gamma}$		
	Fig. 2 c	$R''_{d1} = -R_d(1-n)^2 n \frac{\gamma}{\gamma_{rt}(1-n) + \gamma_b}$		
	Fig. 2 d	$R''_{d2} = R_d(1-n)n^2 \frac{\gamma}{\gamma_{rt}(1-n) + \gamma_b}$		

resistance of the p-n-junction semiconductors, which introduces a negligible error of the same order as the spread of the stabilitron and compensating p-n-junction semiconductor temperature coefficients. In the Table, N is the number of compensating p-n-junction semiconductors: γ'_{rf}, γ'_b and γ'_{p-n} are absolute temperature coefficients in mv/°C of voltages U_{rf}, U_b and that of the p-n-junction semiconductors.

In order to be able to use the formulas given in the Table it is necessary to know γ_{rf}, which can be obtained with the knowledge of U_{rf} from [6,7], as well as γ'_b of the DCA transistor (about 2 mv/°C), n, R_d and γ'_{p-n}. The temperature coefficient of the p-n-junction semiconductors depends on the current they take, which in turn depends on the value of the potential divider R_d and the resistors which shunt the p-n-junction semiconductors.

Figure 4 shows the relation of the current to the absolute-voltage temperature coefficient in mv/°C of the silicon stabilitrons D808-D813 and the germanium diodes D7 and DGTs connected in their forward direction. It will be seen from these curves that for small currents (up to 20 ma) the temperature coefficient variations are of the order of 0.01 $\frac{mv/°C}{ma}$. If more than 2 p-n-junction semiconductors are used for temperature compensation purposes, the variation of the current flowing through them by several dozen milliamperes can provide a total viariation of the temperature coefficient equal to or greater than half the temperature coefficient of one p-n junction. Thus, it is easily possible to control the temperature coefficient of a stabilizer first roughly by withdrawing or adding p-n-junction semiconductors, and then accurately by controlling the current through the p-n junctions.

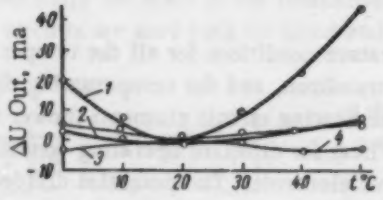

Fig. 5. Variations with temperature of a stabilizer output voltage compensated by different methods.

Accurate adjustments of temperature compensation can also be made conveniently by changing the shunting resistors R'_{d1} and R'_{d2}. If these resistors are not used, the formulas contained in the tables are amended by omitting the term containing them in the denominator under the slanting line. Moreover, the N for such a circuit is determined in advance assuming that R'_{d1} and R'_{d2} are absent. If the value of N thus obtained is fractional, then in using the circuits of Figs. 3a and 3b it is necessary to take the nearest larger integer value of N, add a shunting resistor determined from the Table formulas, and adjust it until the zero value fo the temperature coefficient for the stabilizer is obtained. When the circuits of Figs. 3c and 3d are used it is necessary to selec the nearest smaller integer value of N and proceed in a similar manner until a zero temperature coefficient is obtained.

In the most commonly used circuit (Fig. 1a) for the most widely employed methods of compensation (Fig. 3a and 3c) the values of R'_{d1} and R'_{d2} are determined respectively from

$$R'_{d1} = \frac{N\gamma'_{p-n} R_d (1-n) n}{N\gamma'_{p-n} n + \gamma'_{rf} + \gamma'_b}, \quad (7)$$

$$R'_{d2} = -\frac{N\gamma'_{p-n} R_d (1-n) n}{N\gamma'_{p-n} n + \gamma'_{rf} + \gamma'_b}. \quad (8)$$

For all the remaining circuits these resistors can be determined directly from the Table formulas.

It will be seen from the Table that in using compensation circuits of Figs. 3c and 3d a smaller number of p-n-junction semiconductors is required than for the circuits of Figs. 3a and 3b.

Figure 5 shows the relation between U_{out} and temperature for various practical stabilizer temperature-compensating circuits described by the author of this article in [2]: for U_{out} = 10 v, n = 0.8, R_d = 200 ohm, U_{rf} = 8v, γ'_{rf} = = 4.5 mv/°C and γ'_b = 2 mv/°C. Curves 1 and 2 show respectively the relation of voltage to temperature for temperature compensation by means of MMT thermistors (in the circuit of Fig. 2a the value fo R'_{d1} = 2 ohm, and in Fig. 2c it is R'_{d1} = 770 ohm) and for copper resistors (in the circuit of Fig. 2b it is R'_{d2} = 62 ohm, in Fig. 2d it is R'_{d2} = 410 ohm).

Temperature compensation by means of thermistors can be attained with a accuracy of ± 7 mv only over a narrow temperature range (of the order of ± 10°C to 20°C).

When copper resistors are used it becomes possible to obtain an output voltage stability of the order of ± 6 mv for temperature deviations from 20°C up to 50°C and down to 0°C, thus providing for an output voltage of 10 v a temperature coefficient stabilization of the order of 0.001%/°C. It should be noted that such a stability is only obtained with slow temperature variations. In order to obtain thermal compensation by means of p-n-junction semiconductors in this instance, it is possible to use circuits shown in Figs. 3a and 3c.

Assuming the potential divider current to be 50 ma (U_{out} = 10 v, R_d = 200 ohm) we find for germanium diodes from the curve in Fig. 4 that TC=-1.4 mv/°C. Moreover, we find from the Table for circuits without shunting resistors shown in Figs. 3a and 3c that N = 2.2. If we assume that N = 2 the resulting temperature coefficient without shunting resistors will be +0.25 mv/°C (+0.0025%/°C). By shunting the compensating diodes with a variable resistor as shown in Fig. 3c, the effect of the negative temperature coefficients is raised and it becomes possible to attain a TC stabilization approaching zero. If we assume that N = 3, it is possible to attain a temperature coefficient approaching zero by using a shunting resistance as shown in Fig. 3a.

Curve 3 (Fig. 5) shows the relation between U_{out} of a stabilizer compensated as shown in Fig. 3a, and curve 4 as shown in Fig. 3c.

It will be seen from curves 3 and 4 that in the temperature range of 0-50°C the voltages remain practically constant with a small temperature coefficient, which depends on the accuracy of the compensation adjustment and on other secondary factors.

It is necessary to note the importance of providing similar temperature conditions for all the temperature sensitive elements. For this purpose the reference stabilitron, the DCA transitron, and the compensating diodes must be placed in a separate copper or aluminum box. It is necessary that the heating circuit elements (power transistors, resistors, tubes, transformers, etc.), which radiate different quantities of heat for different operating conditions, should not introduce variations in the heating of the temperature sensitive elements. The potential divider resistors should also be placed away from the sources of heat and in uniform temperature conditions.

CONCLUSIONS

The above computation of the temperature compensation for transistor stabilizers makes it possible to construct devices with highly-stable output voltages. It is best to use for temperature compensation germanium or silicon diodes in a forward connection, using the above-described method for accurate adjustment of temperature compensation.

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RAISED SENSITIVITY OF TUNED CIRCUITS IN INSTRUMENTS WITH REACTIVE TRANSDUCERS

M. L. Shchupak

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pp. 28-33, July, 1961

The modern development of measurement technology is characterized by the application of electrical equipment for measuring practically any physical quantity, including nonelectrical quantities (temperature, humidity, mechanical quantities, etc.).

Instruments whose measuring systems employ tuned circuits are being used increasingly for testing nonelectrical quantities. In these circuits the measured capacitance or inductance forms part of an oscillatory circuit which operates under conditions of resonance or approaching resonance. The variations in capacitance (inductance) lead to variations in the natural resonant frequency of the circuit, thus making it possible to evaluate the measured quantity. Normally the scale of the indicating instrument is calibrated in units of the measured nonelectrical quantity. Tuned circuits are used both for fixed and variable operating frequencies.

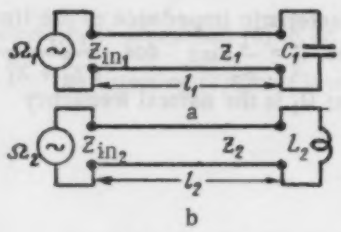

Fig. 1.

In this article we deal with tuned circuits without losses which operate on the fundamental frequency and are formed by reactive transducers only and a transmission line connecting the transducer with the measuring circuit.

Figure 1 shows resonant circuits with reactive transducers, namely, a capacitative transducer (Fig. 1a) and an inductive one (Fig. 1b). The operation of these circuits is based on the variation of the electrical length of the equivalent line due to the variations of the transducer's reactance.

It is known that the resonant frequency of a long line can be represented by the expression

$$\Omega = \frac{\beta l}{\sqrt{L_1 l \cdot C_1 l}}, \qquad (1)$$

where β is the phase constant of the line: L_1 and C_1 are the inductance and capacitance of the line per unit length; l is the length of the line.

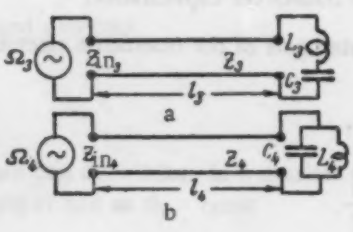

Fig. 2.

It will be seen from (1) that the capacitance and inductance of the line increase with its length, thus determining, as will be shown below, the relatively low sensitivity of resonance lines shown in Fig. 1.

Efforts have been made, for instance in [4], to reduce the above undesirable effect of communication lines on the sensitivity of resonant circuits by using at the output of the lines circuits with lumped constants. Under such conditions the tuned circuits are only used as matching devices, and the existence of the communication line is regarded as an inevitable defect of resonance circuits used for remote measurements.

The sensitivity of resonant-circuit instruments with reactive transducers can be raised by means of the suggested method, which consists of connecting to the output of the communication lines tuned circuits with lumped constants (so-called controlling circuits) with the reactive transducer forming one of the constants. The method is based on the joint utilization of the resonant properties of long lines and tuned circuits with lumped constants.

Figure 2 shows resonant lines with series (Fig. 2a) and parallel (Fig. 2b) tuned circuits. Such circuits are known as resonant controlling circuits.

The sensitivity of these circuits is determined as the ratio of the relative variation of the natural circuit frequency (Ω) to the relative variation of the transducer's inductance or capacitance (T).

For a condition of a fixed operating frequency (Ω_0 = const) the sensitivity of the tuned circuit is

$$S_0=\frac{\left[\frac{d\Omega}{\Omega}\right]_{\substack{\Omega=\Omega_0\\ \omega=\Omega_0}}}{-\frac{dT}{T}}, \tag{2}$$

and for a variable operating frequency ($\Omega = \Omega_0 = \text{var}$) it is

$$S=\frac{\frac{d\Omega}{\Omega}}{-\frac{dT}{T}}, \tag{3}$$

where Ω_0 is the operating frequency; $\omega = 1/\sqrt{LC}$ is the natural frequency of the controlling tuned circuit; L and C are the inductance and capacitance of the controlling tuned circuit.

In these measurements it is assumed that the condition $\Delta T/T \ll 1$ holds.

Resonant circuit with a capacitative transducer. From an examination of the circuit in Fig. 1a we can write:

$$Z_{in1}=-jZ_1 \operatorname{ctg}[\beta_1(l_1-l_1')], \tag{4}$$

where Z_{in1} is the input impedance of the resonant circuit; $Z_1=\sqrt{L_{l1}/C_{l1}}$ is the characteristic impedance of the line; L_{l1} and C_{l1} are the distributed inductance and capacitance; l_1 is the length of the line, $l_1'=\frac{1}{\beta_2}\operatorname{arc}\cos\frac{X_1}{\sqrt{Z_1^2+X_1^2}}$; X_1 is the load of the line; $\beta_1=\Omega_1\sqrt{L_{l1}C_{l1}}=\Omega_1 Z_1 C_{l1}$ is the phase constant of the line; Ω_1 is the natural frequency of the resonance circuit.

According to our assumption we have

$$Z_{in1}=0;\ X_1=-j\frac{1}{\Omega_{01}C_{01}},$$

where Ω_{01} is the basic operating frequency of the resonant circuit; C_{01} is the normal transducer capacitance.

For convenience of analysis both here and subsequently, all the elements forming part of the resonance circuit are denoted by one subscript.

Taking into consideration the above assumptions we obtain from equation (4):

$$\beta_1 l_1=\Omega_{01}Z_1C_{l1}l_1=\arccos\left(-\frac{1}{\sqrt{1+Z_1^2\Omega_{01}^2C_{01}^2}}\right)-\frac{\pi}{2}. \tag{5}$$

Expression (5) is a transcendental equation whose solution by any known method with respect to Ω_{01} will provide the operating frequency of the resonant circuit.

By changing the nominal transducer capacitance the electrical length of the equivalent circuit is also changed according to (4), thus leading to a corresponding variation in the natural frequency Ω_1 of the resonant circuit, which is equal to

$$\Omega_1=\frac{\arccos\left(-\frac{1}{\sqrt{1+Z_1^2\Omega_{01}^2C_1^2}}\right)-\frac{\pi}{2}}{Z_1C_{l1}l_1}. \tag{6}$$

Having differentiated Ω_1 with respect to C_1 according to (2) and having denoted

$$Z_1\Omega_{01}C_{01}=m_1 \tag{7}$$

(known as the equation parameter), we find after appropriate transformations the sensitivity of the resonant circuit with a capacitative transducer at a constant operating frequency

$$S_{01} = \frac{2m_1}{\pi(1+m_1^2)} \quad \text{for } 0 < m_1 < 1. \tag{8}$$

It should be noted that the natural frequency of the resonant circuit in condition Ω_{01} = constant is represented by expression (6) and cannot be measured directly, since it is a function of the reactance of C_1 which in turn depends on the operating frequency Ω_{01} of the circuit.

The above remark applies to the same extent to other resonant circuits with a fixed operating frequency. Nevertheless, this method of analysis is efficient and agrees well with experimental results.

The technique of determining the natural frequencies of resonant circuits for Ω_0 = const is described below (See "Example of a computation and experimental data.")

Under the conditions of a variable operating frequency we have

$$\frac{d\Omega_1}{dC_1} = -\frac{\dfrac{\partial F_1}{\partial C_1}}{\dfrac{\partial F_1}{\partial \Omega_1}}. \tag{9}$$

where F_1 is expression (5) in an implicit form.

By inserting (9) into (3) we find the sensitivity of the resonant circuit for condition $\Omega_1 = \Omega_{01}$ = var:

$$S = \frac{m_1}{m_1 + \beta_1 l_1 (1+m_1^2)}. \tag{10}$$

Resonant circuit with an inductive transducer (Fig. 1b). Considering that

$$Z_{in2} = \infty, \quad X_2 = j\Omega_{02} L_{02} \tag{10a}$$

and denoting

$$\frac{\Omega_{02} L_{02}}{Z_2} = m_2, \tag{11}$$

we can determine, in a manner similar to the above, the sensitivity of the resonant circuit with an inductive transducer and an Ω_0 = const

$$S_{02} = \frac{2m_2}{\pi(1+m_2^2)} \quad \text{for } 0 < m_2 < 1. \tag{12}$$

For a variable operating frequency ($\Omega = \Omega_0$ = var) we have

$$S_2 = \frac{m_2}{m_2 + \beta_2 l_2 (1+m_2^2)}. \tag{13}$$

Resonant line with a controlling series-tuned circuit (Fig. 2a). Considering that

$$Z_{in3} = \infty, \quad X_3 = j\frac{\Omega_{03}^2 L_{03} C_{03} - 1}{\Omega_{03} C_{03}},$$

we can obtain from (4) an equation for the resonant line in Fig. 2a

$$\beta_3 l_3 = \Omega_{03} Z_3 C_{13} l_3 = \arccos \frac{\Omega_{03}^2 L_{03} C_{03} - 1}{\sqrt{Z_3^2 \Omega_{03}^2 C_{03}^2 + (\Omega_{03}^2 L_{03} C_{03} - 1)^2}}. \quad (14)$$

When the controlling circuit is tuned to frequency Ω_{03} we have

$$\Omega_{03}^2 L_{03} C_{03} = 1, \quad (15)$$

and the fundamental operating frequency of the resonant line is determined by the expression

$$\Omega_{03} = \frac{\pi}{2 Z_3 C_{13} l_3}. \quad (16)$$

The detuning of the controlling tuned circuit by a variation in the nominal value of one of its reactance, for instance, of the capacitance, leads to a corresponding variation in the electrical length of the equivalent line and in its natural frequency in a manner similar to that of the two preceding cases

$$\Omega_3 = \frac{\arccos \dfrac{\Omega_{03}^2 L_{03} C_3 - 1}{\sqrt{Z_3^2 \Omega_{03}^2 C_3^2 + (\Omega_{03}^2 L_{03} C_3 - 1)^2}}}{Z_3 C_{13} l_3}. \quad (17)$$

Having differentiated Ω_3 with respect to C_3 according to (2) and denoting

$$\frac{\Omega_3}{Z_3} = m_3, \quad (18)$$

where $\rho_3 = \sqrt{L_{03}/C_{03}}$ is the impedance of the controlling series-tuned circuit, we obtain an expression for the sensitivity of the resonant line with a controlling series-tuned circuit and a capacitative transducer working at a fixed operating frequency

$$S_{03} = \frac{2m_3}{\pi}. \quad (19)$$

For the condition of a variable operating frequency we find the derivative $d\Omega_3/dC_3$ from (9), and by inserting it in (3) we obtain

$$S_3 = \frac{2m_3}{\pi + 4m_3}. \quad (20)$$

It can be shown that the sensitivity of a line with inductive transducers is determined by the same formulas (19) and (20).

Resonant line with a controlling parallel-tuned circuit (Fig. 2b). By analogy with the preceding reasoning and considering that

$$Z_{in4} = 0, \quad X_4 = -j \frac{\Omega_{04} L_{04}}{\Omega_{04}^2 L_{04} C_{04} - 1},$$

we can find the sensitivity of a resonant line with a controlling parallel-tuned circuit and an inductive transducer working at a fixed operating frequency. The sensitivity amounts to

$$S_{04} = \frac{2m_4}{\pi}. \quad (21)$$

where

$$m_4 = \frac{Z_4}{\sqrt{\frac{L_{04}}{C_{04}}}}.$$

The sensitivity of the circuit for a variable operating frequency condition is

$$S_4 = \frac{2m_4}{\pi + 4m_4}. \tag{22}$$

The fundamental operation frequency of the resonant line is determined by expression

$$\Omega_{04} = \frac{\pi}{2Z_4 C_{14} l_4}. \tag{23}$$

It can be shown that the sensitivity of the line with a capacitative transducer is also determined by formulas (21) and (22).

Conclusions of the analysis. It will be seen from formulas (19) and (20) that sensitivity S_{03} is proportional to ρ_3, i. e. it rises with a decrease in the capacitance and an increase in the inductance of the controlling tuned circuit. It follows from the above that resonant lines with controlling series-tuned circuits should have transducers with relatively large capacitative reactances and relatively small inductive reactances.

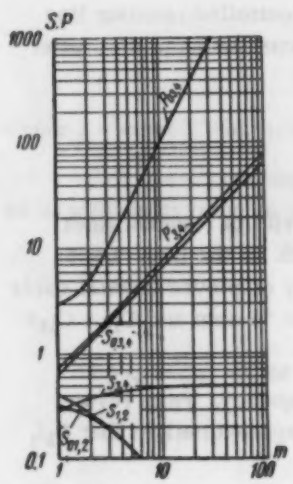

Fig. 3.

By examining on similar lines formulas (21) and (22) we arrive at the conclusion that resonant lines with controlling parallel-tuned circuits should be used with transducers having relatively small inductances and relatively large capacitances.

By comparing S_{03} with S_{01} and S_{04} with S_{02}, we find that sensitivity P_0 (the sensitivity factor) of the controlled resonant lines (Fig. 2) is larger than that of resonant lines (Fig. 1) working with a constant operating frequency, i. e. amplitude modulation (with the remaining conditions being equal):

$$P_{03} = \frac{S_{03}}{S_{01}} = 1 + m_3^2 \tag{24}$$

and

$$P_{04} = \frac{S_{04}}{S_{02}} = 1 + m_4^2. \tag{25}$$

In a similar manner we can find that the sensitivity also rises for a variable operating frequency (frequency modulation).

$$P_3 = \frac{S_3}{S_1} = \frac{2(1 + m_3 + m_3^2 \beta_1 l_1)}{\pi + 4m_3} \tag{26}$$

and

$$P_4 = \frac{S_4}{S_2} = \frac{2(1 + m_4 + m_4^2 \beta_2 l_2)}{\pi + 4m_4}. \tag{27}$$

It follows from (24), (25), (26) and (27) that it is advisable to use controlled resonant lines with a fixed operating frequency when it is mainly required to obtain maximum sensitivity, and to use them with a varying operating frequency when it is required to reduce to a minimum the effect of variations in the transmission line parameters (in particular of temperature) on the resonant circuit characteristics, or when it is required to transmit the measurement results by radio.

In fact, it will be seen from (20) and (22) that if in equality $m_3 = m_4 \gg 1$ is met, the instability of $\beta_3 l_3$ and $\beta_4 l_4$ and of m_3 and m_4 will affect very little the position of the "zero point" and the sensitivity of the controlled resonant lines.

Let us express S_3 and S_4 in terms of S_{03} and S_{04} respectively. By dividing the right-hand sides of formulas (20) and (22) by $\beta_3 l_3$ and $\beta_4 l_4$ respectively we obtain

$$S_3 = \frac{S_{03}}{1 + 2S_{03}}, \quad S_4 = \frac{S_{04}}{1 + 2S_{04}}. \tag{28}$$

It follows from (28) that a rise in the sensitivity of controlled resonant lines with a fixed operating frequency produces a rise in their sensitivity when working with a variable operating frequency, and for a condition of $2S_{03} \gg 1$ and $2S_{04} \gg 1$ we have $S_{3max} = S_{4max} = 0.5$.

From the above formulas it will be seen that a decrease in the length of the communication line produces with the remaining conditions unaltered a decrease in the sensitivity of resonant lines (Fig. 1) and a rise in the sensitivity of controlled resonant lines (Fig. 2).

Figure 3 provides on a logarithmic scale the graphs of functions S, $P = \xi$ (m). Since $m = 1/m_1 = 1/m_2 = m_3 = m_4$ (known as the common controlling parameter) does not depend on the specific values comprising the resonant line circuits, these graphs can be used for computations and as nomograms.

Let us note, incidentally, that the sensitivity of resonant lines which consist of lumped constants only (without a long line) has for any operating condition the limiting value of $S_{max} = 0.5$.

Taking the above reasoning into account it will be seen from (19) and (21) that the controlled resonant line can be regarded as a new type of amplifier, namely, a fundamental frequency (with $\Omega_0 = \text{const}$) detuning multiplier with a detuning factor of

$$M_{03,4} = 2S_{03,4}.$$

Example of computations and experimental data. Let us compute the relative sensitivity of resonant lines shown in Figs. 1a and 2a by taking a given capacitance for the transducer and a given length for the line. Let us assume that $C_{01} = C_{03} + 25 \cdot 10^{-12}$ f. As a communication line let us use the most commonly employed coaxial cable RK-1 with a length of $l_1 = l_3 = 10$ m and with $Z_1 = Z_3 = 75$ ohm and $C_{l1} = C_{l3} = 68 \cdot 10^{-12}$ f/m.

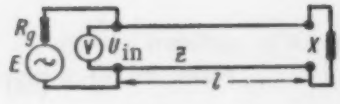

Fig. 4.

We shall begin by determining the operating frequency. From (5) we find for the circuit in Fig. 1a. by means of successive approximations that $\beta_1 l_1 = 1.51$, $F_{01} = \Omega_{01}/2\pi = 4.73 \cdot 10^6$ cps.

From (7) the controlling parameter is $m_1 = 0.0556$.

From (8) and (10) we find by inserting m_1 the sensitivity of the circuit for both conditions: $S_{01} = 0.0355$ and $S_1 = 0.0356$.

According to (16) the operating frequency for the resonant circuit in Fig. 2a is equal to $F_{03} = \Omega_{03}/2\pi = 4.9 \cdot 10^6$ cps.

From (18) we find the controlling parameter $m_3 = 17.3$.

By inserting m_3 in (19) and (20) we find $S_{03} = 11$ and $S_3 = 0.477$.

From (24) and (26) we determine the increase in sensitivity of a controlled resonant line (Fig. 2a) as compared with a simple resonant line (Fig. 1a). We have $P_{03} = 301$ and $P_3 = 13.4$.

The circuit shown in Fig. 4 was used in order to check the theoretical conclusions arrived at in this work. The parameters of the actual circuits used were equal to those given in the above computations.

The measurement technique consisted of the following. The resonant frequency of the line was first determined with the end of the line short-circuited. The resonant frequency was measured by means of an oscillator (GSS-6A) and a tube millivoltmeter (MVL-2M) which was used as a maximum indicator with $R_d = 25$ kohm and $F_{03} = 4.5 \cdot 10^6$ cps.

The operating frequency of the resonant line, whose equivalent circuit is shown in Fig. 1a, was determined by measuring the minimum with R'_d = 13.5 ohm and $F_{01} = 4.3 \cdot 10^6$ cps.

TABLE

Data	F_{01} Mc	S_{01}	S_1	F_{03} Mc	S_{03}	S_3	P_{03}	P_3
Computed	4.73	0.0355	0.0356	4.9	11	0.477	301	13.4
Experimental	4.3	0.0372	0.038	4.5	10.3	0.44	277	11.6

The Q-factor of a long line itself was then determined by the method of detuning to a level of 1.4 for a practical circuit equivalent to the resonant line shown in Fig. 1a, and up to a level of 0.7 for a circuit equivalent to the line shown in Fig. 2a. The Q-factors were calculated by means of the known expression

$$Q = \frac{f_2 + f_1}{2(f_2 - f_1)}$$

and found to be equal to Q_{01} = 21 and Q_{03} = 17.

The input impedance of the line itself (for $X_3 = 0$) was determined from the expression:

$$Z_{in\,03} = \frac{R_d \cdot U_{in\,03}}{E - U_{in\,03}},$$

where E = const is the oscillator voltage; $U_{in\,03}$ is the line input voltage at resonance.

When the controlling tuned circuit $L_{03}C_{03}$ was connected as a load (X) the Q-factor of the resonant line decreased as well as its resonant input impedance Z_{in3} which is determined by the relation

$$Z_{in3} = \frac{R_d \cdot U_{in3}}{E - U_{in3}}.$$

The Q-factor of the controlled resonant line was determined from the expression

$$Q_3 = \frac{Z_{in3} \cdot Q_{03}}{Z_{in\,03}}$$

and was found to be equal to 6.

When capacitance C_{03} = 25 μμf was connected as a load (X) the Q-factor of the resonant line did not change:

$$Q_{01} = Q_1 = 21.$$

Next we determined the sensitivity of a practical circuit equivalent to the line shown in Fig. 1a working at a constant operating frequency. Capacitance C_1 (for Ω_{01} = const) was increased above the nominal value until the voltage across R'_d = 13.5 ohm rose up to a level of 1.4. It was found that:

$$\frac{\Delta C_1}{C_{01}} = 0.64.$$

The relative frequency variation was determined from the well-known equation

$$\frac{\Delta F_{01\,max}}{F_{01}} = \frac{1}{2Q_1} = 0.0238.$$

The sensitivity was determined from (2) as $S_{01} = 0.0372$.

Under the condition of a varying operating frequency, sensitivity (S_1) was determined in the following manner. Capacitance C_1 was varied by ± 10% and the resonance frequencies were determined. It was found that the ratio of the mean values was $\Delta F_1 / F_{01} = 0.0038$. From (3) the sensitivity was obtained as $S_1 = 0.038$.

The values of S_{03} and S_3 were determined in a similar manner and found to be $S_{03} = 10.3$ and $S_3 = 0.44$.

A Q-meter variable capacitor type UK-1 was used as a transducer.

Inductance L_{03} was chosen according to (15).

The calculated and experimental data are shown in a single table.

It will be seen from this Table that the discrepancies between the calculated and experimental data do not exceed the errors of normal design computations.

CONCLUSIONS

The experimental data confirm the theoretical conclusions arrived at in this work and indicate the high efficiency of the suggested method for raising the sensitivity of instruments with resonant circuits and reactive transducers.

The sensitivity obtained by means of controlled resonant circuits is in practice only limited by the effect of parasitic reactances of the controlling tuned circuits.

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PROPERTIES OF CERTAIN ELECTRICAL MEASURING INSTRUMENTS

M. Kh. Shliomovich

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Since January 1, 1961 a new GOST (All-Union State Standard) 1845-59 on "Electrical Measuring Instruments. General Technical Requirements" has come into force. This Standard as distinct from the previously operating general GOST-1845-52 and special standards for various types of instruments (frequency and phase meters, wattmeters, etc.) establishes a single requirement for the effect of all external magnetic fields on the readings of electrical measuring instruments; moreover, this requirement extends to the whole frequency range covered by the instruments (up to 20 kc). The latter requirement is not specified either in any foreign standards or in the recommendations on electrical measuring instruments [1] of the International Electrical Engineering Commission.

The IEEC recommendations, for instance, specify tolerances for external magnetic fields at frequencies only up to 1000 cps for a field strength of 400 av/m (5 oe) in that range. The GOST 1845-59 also specifies the testing of instruments at frequencies up to 1000 cps at a magnetic field strength of 400 av/m. For a further rise in frequency the standard specifies a field strength decreasing in inverse proportion to the frequency.

Such a changing test field strength was first specified at the suggestion of the Kiev "Tochélektropribor" plant in the standard for testing phase meters. However, in further development work and in state testing of various rectifying instruments a considerable effect of magnetic fields on the readings of such instruments was observed. In this connection a study of the rectifying instruments was conducted before compiling GOST 1845-59, in order to establish tolerances for the effect of external magnetic fields on the readings of these instruments.

External magnetic fields produce the following effects on rectifying instruments. The external field induces an emf in the moving coil of the instrument and in the circuit to which the instrument is connected. If the current due to the induced emf is rectified, it produces a deflection of the moving coil from the position it occupied in the absence of an external magnetic field. The rectified current due to the emf induced in the moving coil is large when the instrument reading produced by the measured quantity is small, and it is almost completely independent of the phase relation between the external magnetic field and the measured quantity. The current due to the emf induced in the measuring circuit is superimposed on the current produced by the measured quantity; moreover, the value of the total rectified current which flows in the moving coil and produces its deflection depends on the phase relations between the external magnetic field and the measured quantity.

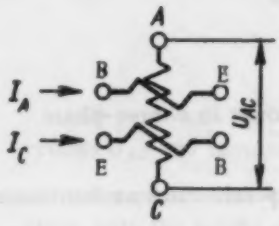

Fig. 1.

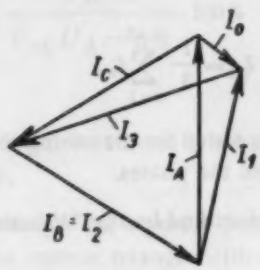

Fig. 2.

B) is the beginning of the winding; E) is the end of the winding.

The testing of instruments has confirmed the above reasoning. When a disconnected rectifying instrument is placed in a magnetic field, its moving coil is deflected from the zero position; this deflection depends on the position of the instrument with respect to the field, and increases with a rising field strength and frequency. This effect is produced by the rectified current due to the emf induced in the moving coil of the intrument.

If a rectifying instrument with its magnetic coil registering a large deflection due to a measured quantity is placed in a magnetic field whose phase varies periodically with respect to the phase of the measured quantity, the instrument readings will also vary periodically. Moreover, the pointer will oscillate about a position corresponding to the measured value, and the amplitude of its oscillation will increase if the external field rises in intensity or frequency. This effect is due to the emf induced in the measuring circuit.

If an instrument with a small deflection of its moving coil (for instance, at the beginning of the scale) is placed in a magnetic field whose phase varies periodically, the moving coil will oscillate about a position corresponding to a larger deflection than that due to the measured quantity. This effect is produced by the combined effect of the emfs induced in the moving coil and measuring circuit to which the instrument is connected.

In the case of rectifying instrument types which do not meet the requirements with respect to external magnetic fields specified for all the instruments of other systems, it was found advisable on the basis of the above investigations to specify a maximum external magnetic field frequency f_k up to which these instruments meet the general requirements. In this case the field strength for the frequencies up to f_k inclusive should be equal to 400 av/m (5 oe), and for higher frequencies f the field strength should be determined from the formula $H = 400\ f_k/f$ av/m. In such a case the Standard provides for an appropriate marking of the scale with the frequency f_k.

Thus, conforming to the inherent defect of rectifying instruments, which consists of their readings being affected by high-frequency external fields, the GOST 1845-59 provides less stringent requirements for these instruments. The same applies to electronic instruments whose readings are also affected to a considerable extent by external magnetic fields. In using such instruments it is always advisable to note their markings, and if necessary allow for a larger measurement error due to the effect of external fields on the instrument. It should be noted that in using these instruments it is in practice very easy to reduce considerably or even completely eliminate the effect of external magnetic fields. For this purpose it is sufficient to find a position for a disconnected instrument in which its pointer deviates from the zero position as little as possible.

From January 1, 1961, the new GOST 8476-60, which determines the technical requirements for wattmeters and VAR meters has also come into effect. As distinct from the previously operating GOST 1845-52 and GOST 8476-57 the new Standard provides considerably more stringent requirements for the testing conditions of three-phase

wattmeters intended for measuring power in uniformly loaded phases. In checking such instruments the new GOST tolerates, contrary to the old GOST 1845-59, asymmetry in three-phase voltages or currents amounting to 1% only, despite the fact that in practice three-phase voltages (currents) are considered to be symmetrical if their asymmetry does not exceed 5% [2,3].

According to GOST 1845-59, the asymmetry of three-phase currents (or voltages) is determined by the expression

$$\sqrt{\frac{2}{3}\sum_{\kappa=1}^{\kappa=3}\delta_\kappa^2}\,,$$

where

$$\delta_\kappa=\frac{I_\kappa-I_m}{I_m} \quad \text{and} \quad I_m=\frac{1}{3}\sum_{\kappa=1}^{\kappa=3}I_\kappa.$$

As an example let us examine two wattmeter circuits intended for measuring the power in a three-phase three-conductor network with a load evenly distributed between the phases.

Figure 1 shows the circuit of wattmeter D162 (D172) manufactured by the "Vibrator" plant. Such an instrument reads power P_a which is equal to

$$P_a=\dot{U}_{AC}(\dot{I}_A-\dot{I}_C). \tag{1}$$

It is known that the actual power P_{aa} of a three-phase three-conductor network [3] is equal to:

$$P_{aa}=\dot{U}_{AB}\dot{I}_A+\dot{U}_{CB}\dot{I}_C. \tag{2}$$

It can easily be shown that in the case of a strictly symmetrical load the wattmeter readings will correspond to the actual power in the network, which can be written as:

$$P_{as}=P_{aas} \tag{3}$$

Any deviation from symmetrical loading produces a circuit error which can be shown in the following manner.

An asymmetrical current system can be obtained from a symmetrical one by supplementing two of the existing currents with currents $\dot{I}_0$ and $-\dot{I}_0$*). In a particular case (Fig. 2) let

$$\begin{aligned}\dot{I}_1&=\dot{I}_A+\dot{I}_0,\\ \dot{I}_2&=\dot{I}_B,\\ \dot{I}_3&=\dot{I}_C-\dot{I}_0,\end{aligned} \tag{4}$$

where $\dot{I}_1$, $\dot{I}_2$ and $\dot{I}_3$ are the currents in phases A, B anc C for an asymmetrical load.

The wattmeter will then read power P_a which is equal to

$$P_a=\dot{U}_{AC}(\dot{I}_1-\dot{I}_3)=\dot{U}_{AC}(\dot{I}_A-\dot{I}_C+2\dot{I}_0)=P_{as}+2\dot{U}_{AC}\dot{I}_0, \tag{5}$$

whereas the power in the network is

$$P_{aa}=\dot{U}_{AB}\dot{I}_1+\dot{U}_{CB}\dot{I}_3=\dot{U}_{AB}(\dot{I}_A+\dot{I}_0)+\dot{U}_{CB}(\dot{I}_C-\dot{I}_0)=P_{aas}+\dot{U}_{AC}\dot{I}_0. \tag{6}$$

*This solution has been suggested by prof. M. L. Levin, D. Sc.

The relative circuit error is

$$\gamma_{sct}=\frac{P_a-P_{aa}}{P_{aa}}\cdot 100\%. \tag{7}$$

$P_{aa} \approx P_{aas} = P_{as}$ for a small I_0.

Hence, from (5), (6) and (1) we obtain

$$\gamma_{sct}=\frac{\dot{U}_{AC}\cdot\dot{I}_0}{\dot{U}_{AC}(\dot{I}_A-\dot{I}_C)}\cdot 100\% = \frac{I_0}{|\dot{I}_A-\dot{I}_C|}\cdot\frac{\cos(\dot{U}_{AC}^{\wedge}\dot{I}_0)}{\cos[\dot{U}_{AC}^{\wedge}(\dot{I}_A-\dot{I}_C)]}\cdot 100\%. \tag{8}$$

It is obvious that for a fixed value of I_0 we shall obtain the maximum γ_{sct} when the directions of the two vectors $\dot{U}_{AC}$ and $\dot{I}_0$ coincide.

Let us examine an example for a purely resistive load. In this case the direction of vector $\dot{I}_0$ (Fig. 3a) coincides with the height of the current triangle with respect to side $\dot{I}_2 = \dot{I}_b$ (Fig. 3b).

If $I_A = I_B = I_C = I_H$, and $I_0 = mI_H$, we have

$$I_1=I_2=\sqrt{\left(\frac{1}{2}I_H\right)^2+\left(\frac{\sqrt{3}}{2}I_H+mI_H\right)^2}=(1+n)I_H,$$

$$I_m=\left(1+\frac{2}{3}n\right)I_H, \tag{8a}$$

$$|\delta_1|=|\delta_2|=\frac{1}{3}n,\quad |\delta_3|=\frac{2}{3}n$$

and for an asymmetry up to 5% we obtain

$$n<\frac{0.05}{\sqrt{\frac{2}{3}\left(\frac{1}{3^2}+\frac{1}{3^2}+\frac{2^2}{3^2}\right)}}=0.075. \tag{8b}$$

For a maximum permissible n = 0.075 we have m = 0.087, i. e. $I_0 = 0.087\ I_H$.

Since for $\cos\varphi = 1$ we have

$$\cos[\dot{U}_{AC}^{\wedge}(\dot{I}_A-\dot{I}_C)]=1 \tag{8c}$$

and $|\dot{I}_A - \dot{I}_C| = \sqrt{3}\ I_H$, we obtain from (8) that $\gamma_{sct} = 5\%$.

Thus, in the above example with currents asymmetrical within the tolerances of a load which is in practice considered to be symmetrical, the wattmeter error amounts to 5%. This means that if a strictly symmetrical load is replaced by a load considered to be asymmetrical in practice, the readings of the wattmeter made according to Fig. 1 may change by as much as 5%.

Another possible circuit consists of a single-phase wattmeter which measures power in a symmetrical three-phase network, or of a single-element wattmeter with an artificial zero point [3].

With an appropriate calibration the wattmeter will read in this circuit power P_a, which is equal (if the wattmeter measuring element is connected to the first phase) to:

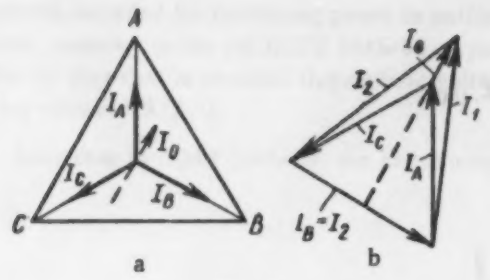

Fig. 3.

$$P_a = 3\,\dot{U}_{A0}\,\dot{I}_A. \tag{9}$$

Providing the load is strictly symmetrical the readings of such an instrument correspond according to [3] to the power in the network, i. e. they satisfy equation (3).

The largest circuit error of such a wattmeter will occur when $\dot{I}_1 = \dot{I}_A + \dot{I}_0$ and the direction of vector $\dot{I}_0$ coincides with the direction of vector $\dot{I}_A$ (Fig. 4a and b).

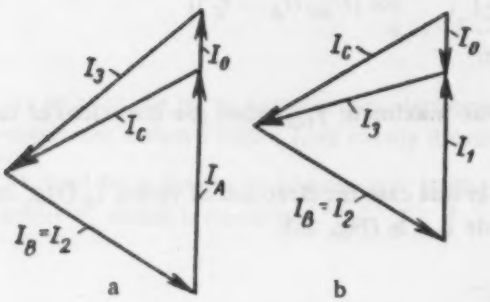

Fig. 4.

Moreover, $I_1 = I_A \pm I_0 = (1 \pm m) I_H$,

$$I_3 = I_B = I_H,$$

$$I_3 = \sqrt{I_C^2 + I_0^2 \pm 2 I_C I_0 \cos\frac{\pi}{3}} =$$

$$= \sqrt{1 \pm m + m^2}\; I_H \approx \left(1 \pm \frac{1}{2} m\right) I_H,$$

$$I_{cp} = \left(1 \pm \frac{1}{2} m\right) I_H,$$

$$|\delta_1| = |\delta_2| = \frac{1}{2} m, \quad |\delta_3| = 0$$

and for an asymmetry of 5% we have

$$m < \frac{0.05}{\sqrt{\frac{2}{3}\left(\frac{1}{2^2} + \frac{1}{2^2} + 0\right)}} = 0.087,$$

i. e. we obtain the maximum possible value of $I_0 = 0.087\ I_H$.

It can easily be seen that in the above case the circuit error is equal to

$$\gamma_{sct} = \frac{I_0}{I_A} \cdot 100\% \tag{10}$$

i. e. $\gamma_{sct} = 8.7\%$.

This means that if a strictly symmetrical load is replaced by a load considered in practice to be symmetrical, the readings of such a wattmeter may change by a much as 8.7%.

The above two examples show that such wattmeters can only be used for measuring power in three-phase networks when the imbalance in the loading of phases produces a very small asymmetry in the currents or voltages, amounting to 1-2%, i. e. only with an extremely uniform load. For a larger asymmetry unacceptably large errors may occur in the readings of the above instruments. This circumstance should be taken into account in selecting instruments for measurements in any given network. The scales of these instruments should be marked as specified in GOST 1845-59 by an exclamation mark, thus drawing the attention of the users to these instruments' specific properties which should be indicated according to the Standard in the instructions or certificates accompanying them.

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UTILIZATION OF THE HALL EFFECT FOR INVESTIGATING THE PROPERTIES OF FERROMAGNETIC CORES

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There is a large amount of known research into the ac magnetic characteristics of closed circuit samples, and samples with air gaps. Such investigations are usually carried out by means of an oscilloscope and a ferrometer [1]. However, until recently there have been no convenient instruments for direct measurements of induction in the air gap.

New possibilities in this connection are provided by the Hall effect, whose essence is described, for instance, in [2]. It has been suggested in [3,4,5] to use the Hall effect for plotting dynamic magnetization curves of ferromagnetic materials. It has also been suggested that Hall emf transducers made from the ferromagnetic materials under investigation should be used. However, due to the small value of the Hall constant for ferromagnetic materials, it has been found impossible to obtain the required sensitivity.

The use for this purpose of special semiconductor Hall emf transducers is, therefore, of considerable interest. The sensitivity of these transducers owing to their large Hall constant is several orders higher than that of ferromagnetic material transducers.

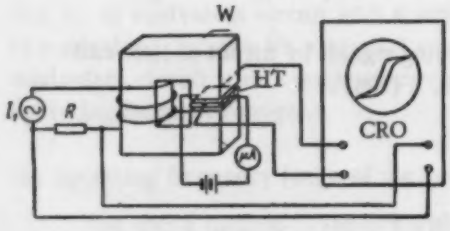

Fig. 1.

The semiconductor Hall emf transducer is placed in the air gap of the core made of the material under investigation (Fig. 1).

The relation between the Hall emf at the transducer output to the current in the magnetizing winding of the core will represent faithfully the relation between induction B, or the intensity of magnetization, and the magnetic field strength H. This relationship can be observed on a cathode-ray oscilloscope. Moreover, the air gap provided for the transducer must not affect the magnetic flux; in other words, the reluctance of the core must be considerably larger than that of the air gap. This method can be used for investigating ferromagnetic materials which have a core with an air gap into which the transducer can be placed. This method is especially convenient for analyzing the behavior of cores at various frequencies and selecting them for linear converters which use Hall emf transducers.

However, it is generally difficult to make cores of a special shape. For samples of a simpler geometrical shape the device shown in Fig. 2 can be used. Section 1 of the core is made of a magnetic material with high permeability and a high saturation induction. Section 2 consists of the tested sample of a simple shape. The shape of the main magnetic core cross section and that of the tested material sample must be the same. The total reluctance of the sample. The reluctance of the remaining parts must be sufficiently small. The better this condition is fulfilled, the higher will be the accuracy of measurements. For precision measurements it is possible to place film transducers on the tested sample. It is then advisable to use film transducers which provide a considerable reduction in the size of the air gap.

It is also advisable to note other sources of errors in this circuit due to the Hall emf transducers themselves; namely:

inaccuracies in positioning Hall electrodes with respect to the equipotential line, thus making the Hall emf distinct from zero, when B = 0. The compensation of this effect is dealt with in [6];

the effect of temperature on the Hall emf;

the effect of induction B on the transducer's resistance;

the presence of stray ac emfs in the Hall electrode circuit.

By means of a rational design of the circuit and the required compensations [6] it is possible to reduce the total error of this method to the same order of magnitude as exists in similar instruments based on other principles.

the simplicity of its design can be preserved at the same time.

The above method was used for selecting cores of various materials for linear converters using Hall emf transducers. Samples made of transformer steel, permalloy and ferrites were tested. The oscillogram (Fig. 1) shows one of the dynamic magnetization curves obtained by this method.

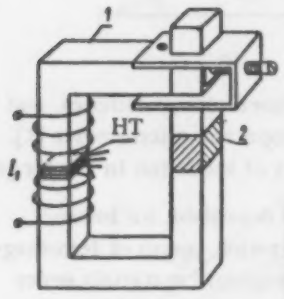

Fig. 2.

Owing to the fact that Hall emf transducers operate practically instantaneously, this method can be applied for determining magnetization curves at very high frequencies, which is especially important in many cases.

By using low-frequency oscillators with appropriate oscilloscopes it becomes possible to obtain by means of this method magnetic characteristics approaching the static condition. Thus, the characteristics obtained by the oscillographic method with a Hall emf transducer and a 15 cps signal differed only by 1-2% from the static characteristics obtained with the same transducer.

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INDUCTANCES, LOSS RESISTANCES AND Q-FACTORS OF INDUCTION COILS

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Any actual coil possesses, in addition to inductance, resistance and self-capacitance. It can be assumed that an actual coil can be represented with sufficient accuracy [1] by means of an equivalent circuit (Fig. 1) consisting of the following elements:

L_c - the effective coil inductance; R_c - the effective coil resistance; C_c - the self-capacitance of the coil, this capacitance can be considered lumped and independent of frequency up to frequencies of $1/3\, f_0$ [3,5], where f_0 is the natural frequency of the coil; G - the coil conductance equivalent to the dielectric losses in the capacitance C_c; $Q = L_c/R_c$ - is the effective Q-factor of the coil.

If the two-terminal network shown in Fig. 1a is replaced by that of Fig. 1b and the value of G is included in that of resistance R_c [1, 2, 3], it becomes possible to represent the impedance of the network by the expression

$$R+j\omega L=\frac{R_c+j\omega L_c\left(1-\omega^2 L_cC_c-\frac{C_c}{L_c}R_c^2\right)^2}{(1-\omega^2 L_cC)^2+\omega^2 R_c^2C_c^2}. \qquad (1)$$

For frequencies lower than the natural frequency f_0 of the coil and for values of Q_C which are not too small, it is possible to obtain from (1) the following expressions:

$$L=\frac{L_c}{1-\omega^2 L_c C_c}\approx\frac{L_c}{1-(f/f_0)^2}, \tag{2}$$

$$R=\frac{R_c}{(1-\omega^2 L_c C_c)^2}\approx\frac{R_c}{[1-(f/f_0)^2]^2}, \tag{3}$$

$$Q=\frac{\omega L}{R}=Q_c(1-\omega^2 L_c C_c)=Q_c[1-(f/f_0)^2]. \tag{4}$$

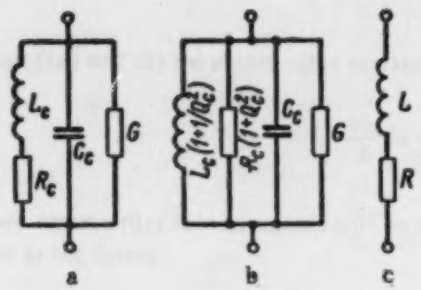

Fig. 1. a) equivalent circuit with a series connection; b) equivalent circuit with a parallel connection; c) equivalent circuit with a series connection for the operating frequency range.

The values of L, R, and Q can be measured by means of existing measuring instruments, and those of L_C, R_C and Q_C can be calculated from (2), (3) and (4), providing that the natural resonant frequency of the coil is known. The main disadvantage of this method consists in L_C and especially C_C of many coils becoming affected by frequency when it approaches the value of f_0 [3,5,7,8], thus invalidating the equality

$$f_0=\frac{1}{2\pi\sqrt{L_c C_c}}, \tag{5}$$

which is incorporated in (2), (3) and (4), and making it impossible to correct accurately the above expressions in the operating frequency range of the coil when the measured value of f_0 is used in them.

The above method is applied with certain modifications in Q-meters whose variable capacitor C_0 is small, and provide the tuning of certain measured coils with very low values of C_0.

The measured values of the Q-factor are corrected by means of the expression

$$1-k^2=1-(f/f_0)^2=1-\frac{C_c}{C_0+C_c}=\frac{C_0}{C_0+C_c}, \tag{6}$$

provided in [6,7,8] and derived from (5) and

$$f=\frac{1}{2\pi\sqrt{L_c(C_0+C_c)}}.$$

The above method of determining $(1-k^2)$ is only used in Q-meters.

The correction factor $1-k^2=1-\omega^2 L_C C_C$ can also be calculated [1,5] from the values of L_C and C_C obtained graphically (Fig. 2). In view of the fact that in the majority of coils measurements must be limited up to a frequency $\leq 1/3\, f_0$, the graphical method has the following desadvantages: the straight line $1/f^2=\varphi(C_g)$ is plotted with a considerable error, since there are no data for its left-hand part; with a turning capacitor value of $Cg \geq 8 C_C$ small variations of ΔC_g (of the order of C_C) cannot provide accurate indications of resonance, thus making the readings of the above values inaccurate; the C_C scale becomes coarse and hence its readings inaccurate; the graphical determination involves additional operations and calculations introducing additional errors; the accuracy in determining L_C and C_C depends on double the relative error in determining frequency and on the error of the tuning capacitor calibration.

We given below a more accurate and convenient method of determining L_C, R_C and Q_C, based on measuring the inductance of the coil in the operating frequency range. For this purpose any measuring instrument suitable

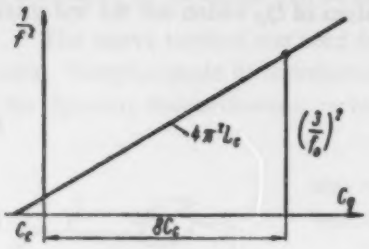

Fig. 2.

for a given coil can be used. The above method is derived from (2) which can be written in the form:

$$\frac{1}{L} = \frac{1}{L_c} - \omega^2 C_c. \tag{7}$$

If the inductance of the coil is measured at frequencies f_1 and f_2 we have

$$\frac{1}{L_1} = \frac{1}{L_c} - \omega_1^2 C_c, \tag{8}$$

$$\frac{1}{L_2} = \frac{1}{L_c} - \omega^2 C_c. \tag{9}$$

from (8) we obtain

$$C_c = \frac{\frac{1}{L_c} - \frac{1}{L_1}}{\omega_1^2}, \tag{10}$$

and from (8) and (9) we have

$$C_c = \frac{\frac{1}{L_1} - \frac{1}{L_2}}{\omega_2^2 - \omega_1^2}. \tag{11}$$

Since the left-hand sides of equations (10) and (11) are equal, we can equate their right-hand sides and thus obtain after certain transformations the expression

$$L_c = L_1 L_2 = \frac{(n^2-1)}{n^2 L_2 - L_1}, \tag{12}$$

where $n = \omega_2 / \omega_1$.

In order to determine the values of R_C and Q_C it is necessary to know the value of the correcting factor $1 - k^2 = 1 - \omega^2 L_C C_C$. We can obtain from (2) and (12) the following values for $1 - k^2$: for frequency f_1

$$1 - k_1^2 = \frac{(n^2-1) L_2}{n^2 L_2 - L_1}, \tag{13}$$

for frequency f_2

$$1 - k_2^2 = \frac{(n^2-1) L_1}{n^2 L_2 - L_1}. \tag{13a}$$

By using expressions (13) and (13a) we can write equations (3) and (4) in the following manner: for frequency f_1

$$R_{c1} = R_1 \left[\frac{(n^2-1) L_2}{n^2 L_2 - L_1}\right]^2, \tag{14}$$

$$Q_{c1} = Q_1 \frac{n^2 L_2 - L_1}{(n^2-1) L_2}; \tag{15}$$

for frequency f_2

$$R_{c2} = R_2 \left[\frac{(n^2-1) L_1}{n^2 L_2 - L_1}\right]^2, \tag{14a}$$

$$Q_{c2} = Q_2 \frac{n^2 L_2 - L_1}{(n^2-1)\, L_1}, \tag{15a}$$

where $n = f_2/f_1$.

It should be noted that in deriving (7), (8) and (9) we have not taken into account the effect of resistance R on the values of L_1 and L_2, which leads to a certain error in calculating $1 - k^2$. The value of this error $\Delta L/L$ can be determined from (1) and (2).

From (1) we obtain

$$L' = \frac{L_c\left(1 - \omega^2 L_c C_c - \frac{C_c}{L_c} R_c^2\right)}{(1-\omega^2 L_c C_c)^2 - \omega^2 R_c^2 C_c^2}. \tag{1a}$$

From (1a) and (2) we obtain after certain transformations

$$\frac{\Delta L}{L} = \frac{1}{Q_c^2 [(f_0/f)^2 + (f/f_0)^2 - 2] + (f/f_0)^2}. \tag{16}$$

If we assume that measurements will be made at frequencies not exceeding $1/3\, f_0$, expression (16) can also be written in the form:

$$\frac{\Delta L}{L} \approx \frac{1}{7Q_c^2}. \tag{16a}$$

Thus the error $\Delta L/L$ due to neglecting the effect of R_C on the measured value of the inductance is very small, and for coils with a Q-factor exceeding 10 it is negligible.

The accuracy in determining the correction term $1 - k^2$ in (13) and (13a) is also affected by errors in setting the frequencies f_1 and f_2 in order to obtain the ratio $n = f_2/f_1$. However, when relative errors in frequencies $\Delta f_1/f_1$ and $\Delta f_2/F_2$ are approximately equal, they will not affect the accuracy of $1 - k^2$.

The choice of the ratio $n = f_2/f_1$ is determined by the following considerations: n should be large in order to obtain values of L_1 and L_2 convenient for measurement purposes; yet for coils with ferromagnetic cores, n must not be too large so as to prevent permittivity being affected by frequency. Therefore, the value of n for each particular case should be selected according to the parameters of the tested coil. For the overwhelming majority of cases it is convenient to have $n = 2$.

It is also necessary to determine the degree to which the various errors in measuring L_1 and L_2 affect the accuracy in determining $1 - k^2$. Let L_1 and L_2 be measured with relative error of $\Delta L_1/L_1$ and $\Delta L_2/L_2$ respectively, i.e. let

$$L_1' = L_1\left(1 + \frac{\Delta L_1}{L_1}\right); \quad L_2' = L_2\left(1 + \frac{\Delta L_2}{L_2}\right).$$

If we denote

$$\delta_1 = \frac{1 + \frac{\Delta L_1}{L_1}}{1 + \frac{\Delta L_2}{L_2}}; \quad \delta_2 = \frac{1 + \frac{\Delta L_2}{L_2}}{1 + \frac{\Delta L_1}{L_1}},$$

expression $1 - k^2$ will assume the form

$$(1-k_1^2)' = \frac{(n^2-1)\, L_2}{n^2 L_2 - \delta_1 L_1}. \tag{17}$$

From (12) and (17) we obtain

$$\frac{\Delta(1-k_1^2)}{1-k_1^2} = \frac{\dfrac{(n^2-1)L_2}{n^2L_2-\delta_1L_1} - \dfrac{(n^2-1)L_2}{n^2L_2-L_1}}{\dfrac{(n^2-1)L_2}{n^2L_2-L_1}},$$

whence after certain transformations we obtain

$$\frac{\Delta(1-k_1^2)}{1-k_1^2} = \frac{\delta_1-1}{n^2L_2/L_1-\delta_1}.$$

Since the values of L_1 and L_2 are very close to each other, it is possible to assume that $L_2/L_1 \approx 1$ and by neglecting the value of $(\Delta L_1/L_1 - \Delta L_2/L_2)$ with respect to $(n^2 - 1)$ we finally obtain:

$$\frac{\Delta(1-k_1^2)}{1-k_1^2} < \frac{\Delta L_1/L_1-\Delta L_2/L_2}{n^2-1}. \tag{18}$$

By means of similar transformation we can obtain for $1 - k_2^2$ the following expression:

$$\frac{\Delta(1-k_2^2)}{1-k_2^2} < \frac{n^2}{n^2-1}\left(\frac{\Delta L_1}{L_1}-\frac{\Delta L_2}{L_2}\right). \tag{19}$$

In practice the values of L_1 and L_2 are very close to each other and are measured in a relatively narrow frequency range by means of the same inductance measuring device. Hence, $\Delta L_1/L_1 \approx \Delta L_2/L_2$ and the relative error in determining L_1 and L_2 due to the inductance measuring device will not in practice affect the value of $1 - k^2$.

Let us give an example in determining L_C and $(1 - k^2)$ for a toroidal coil with a ferrimagnetic core.

The measured data are: $L_1 = 100.5$ mh for $f_1 = 10$ kc; $L_2 = 105.3$ mh for $f_2 = 15$ kc; the effective inductance value measured at very low frequences is $L_C \approx 97$ mh for $f_0 = 52$ cps; the additional Q-meter capacitor for bringing the circuit to resonance at 15 kc amounts to 1055 $\mu\mu$f; the self-capacitance of the coil $C_C = 96 \pm 2$ $\mu\mu$f; and the coil Q-factor is $Q \approx 50$.

By using these data for calculations: 1) from (6) we have

$$1-k_2^2 = 0.916 \pm 0.002,$$

whence we obtain from (2)

$$L_c = L_2(1-k_2^2) = 96.45 \pm 0.3 \text{ mh};$$

2) from expression (11) we have $L_C \approx 96.94$ mh; 3) from (13) and (13a) we obtain for the correction terms the following values: for $f_1 = 10$ kc we have $1 - k_1^2 = 0.965$; for $f_2 = 15$ kc we have $1 - k_2^2 = 0.921$; 4) the values of the correction terms calculated from the natural resonant frequency of the coil are for $f_1 = 10$ kc we have $1 - k_1^2 = 1 - (f_1/f_0)^2 = 0.963$; for $f_2 = 15$ kc we have $1 - k_2^2 = 1 - (f_2/f_0)^2 = 0.9165$.

It will be seen from the above results that the values of L_C and $(1 - k^2)$ determined by different methods are practically equal. It will also be seen that in the case of a toroidal coils operating at low frequencies the method of determining L_C, R_C and Q_C by measuring the inductance is preferable to other methods, since only one additional measurement of the inductance is needed. It is known that the determination of the natural resonant frequency [8] or the self capacitance [4,9] of the coil involves much more complicated measurements and additional measuring equipment. Hence, the new method is always preferable providing suitable induction measuring devices are available.

CONCLUSIONS

The suggested new method of determining L_C, R_C and Q_C is applicable to all direct and indirect methods of

measuring inductances. Since the values of the latter are very close to each other and are determined in a relatively narrow frequency band by means of the same measuring device, their relative errors $\Delta L_1 / L_1$ and $\Delta L_2 / L_2$ are of the same order and according to (18) and (19) will not have an appreciable effect on the value of $1 - k^2$.

For coils with cores whose permeability varies with frequency the ratio $n = f_2 / f_1$ should be selected in such a way that in the measured frequency range the permeability should remain practically unchanged. For coils with cores whose permeability does not vary with frequency, n can be chosen so as to obtain values of L_1 and L_2 suitable for measuring purposes.

The new method requires only one additional measurement of the coil inductance, and is simpler than any other method. Providing all the above mentioned conditions are observed and a measuring generator whose $\Delta f_1 / f_1 \approx \approx \Delta f_2 / f_2$ is used, this method can provide a far more precise determination of L_C, R_C and Q_C.

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DIRECT-READING COMPARATOR CAPACITANCE BRIDGES

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Direct-reading comparator bridges DRCB and precision comparator bridges PCB are intended for measuring the deviations of capacity from a nominal value, as well as measuring the difference in loss angles of the tested and reference capacitors.

The bridges are designed to be used with an external reference capacitor and can be used in a balanced or unbalanced condition.

In the latter case the tuning indicator provides a direct reading of the capacitance difference in percentages, thus greatly increasing the productivity of labour in using this bridge.

The technical characteristics of the DRCB and PCB instruments are given in the Table below.

The bridges of both types are fed from an external stabilized 1000 cps oscillator with a variable voltage of 0 to 20 v.

The bridge tuning indicator consists of a Tesla TM-622 cathode-ray tube. With a slight modification the Soviet-made indicator ÉLUR-2 can also be used.

Characteristics	DRCB	PCB
Measurement range of capacity deviations from the normal, %	0- ± 20	0- ± 10
Measurement range of loss angles, %	0- 1.8	0 - 0.2
Calibration of the capacity deviation scale, %	0.05	0.005
Range of the measured cpacitances, μμf;	From 5000 up	
Calibration of the loss angle scale, %	0.1	0.01
Maximum error in measuring capacity deviations, %		
by the balanced bridge method	± 0.1	± 0.02
by the unbalanced bridge method	± 0.5 (for the range of ± 5%)	± 0.02 (for the range of ± 0.5%)
	± 1.5 (for the range of ± 10 and ± 20 %)	± 0.2 (for the range of ± 10%)
Error in measuring the loss angle, %	± 0.1	± 0.01

The instruments are intended for working at an ambient temperature of 15-35°C and a relative humidity not exceeding 80 %.

The schematic of the DRCB instrument is given in Fig. 1 and that of the PCB instrument in Fig. 2.

The circuits differ in the construction of arm Z_2. The additional capacitors in the arm Z_2 of the DRCB circuit serve to extend the loss angle measurement range.

The comparator bridges have two built-in arms Z_1 and Z_2, the other two arms, Z_X and Z_0, consist of the compared (measured C_X and reference C_0) capacitors. The bridge is provided with special sockets for plugging in these capacitors.

Arm Z_2 serves to vary continuously the arms ratio of the bridge. In the CRCB instrument it consists of a fixed wire-wound resistor, a set of switched wire-wound resistors, and a slidewire with a continuously variable range from 0 to 1 ohm. The phase is compensated by means of a variable air capacitor C_2. The loss angle range can be extended by connecting fixed capacitors C_2' and C_2''. The effective values of these capacitors are adjusted when the loss angle scale is calibrated.

Arm Z_1 in both bridges consists of a fixed wire-wound resistor placed inside a screen and shunted by a semivariable capacitor C_1. This capacitor serves to set the zero in calibrating the loss angle scale.

The reactive component in both the DRCB and PCB type bridge is balanced by varying the ratio of resistors R_1 and R_2, and the resistive component by the variable capacitor C_2 and in the case of the DRCB instrument also by the additional capacitors C_2' and C_2''.

In order to examine the balance conditions of the bridge it is possible to represent circuits given in Figs. 1 and 2 by a single simplified circuit (Fig. 3). Capacitors C_X and C_0 in the arms Z_h and Z_0 are assumed to be ideal; ohmic resistances connected in series with them represent their losses. When the ac bridge is balanced we have

$$Z_1 Z_0 = Z_2 Z_x, \tag{1}$$

where

$$Z_1 = \frac{1}{\frac{1}{R_1} + j\omega C_1}; \qquad Z_0 = R_0 - \frac{j}{\omega C_0};$$

$$Z_2 = \frac{1}{\frac{1}{R_2} + j\omega C_2}; \qquad Z_x = R_x - \frac{j}{\omega C_x};$$

let

$$\operatorname{tg}\delta_1 = \frac{1}{\omega C_1 R_1};$$

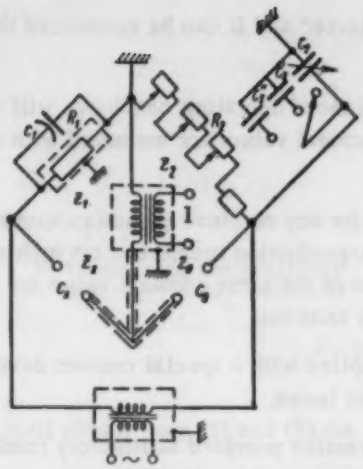

Fig. 1. R_1 - 1000 ohm; C_1 - 25-150 $\mu\mu f$; R_2 - 800 + 2 · 100 + 2 · 10 + 1 ohm; C_2 - 20-1000 $\mu\mu f$; C_2' - 1000 $\mu\mu f$; C_2'' - 2000 $\mu\mu f$; I - indicator.

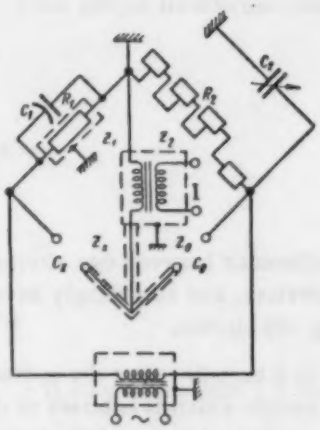

Fig. 2. R_1 - 1000 ohm; C_1 - 25-150 $\mu\mu f$; R_2 - 900 + 18 · 10 + 21 · 1 + 22 · · 0.05 ohm; C_2 - 18 - 510 $\mu\mu f$; I - indicator.

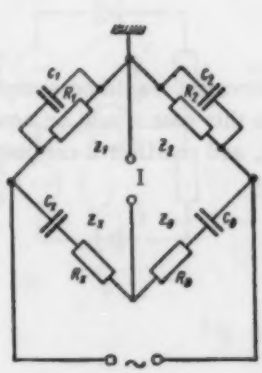

Fig. 3.

$$\operatorname{tg}\delta_2 = \frac{1}{\omega C_2 R_2};$$

$$\operatorname{tg}\delta_x = \omega C_x R_x;$$

$$\operatorname{tg}\delta_o = \omega C_o R_o.$$

By inserting these values in (1) and after several mathematical operations we equate the real and imaginary parts.

Then we obtain for

a) $$\frac{1}{C_x} = \frac{R_1}{C_o R_2}\left(\frac{1}{\operatorname{tg}\delta_1 \operatorname{tg}\delta_2} + \frac{\operatorname{tg}\delta_o}{\operatorname{tg}\delta_1} - \frac{\operatorname{tg}\delta_o}{\operatorname{tg}\delta_2} + 1\right).$$

In our circuits

$$\frac{1}{\operatorname{tg}\delta_1 \operatorname{tg}\delta_2} + \frac{\operatorname{tg}\delta_o}{\operatorname{tg}\delta_1} - \frac{\operatorname{tg}\delta_o}{\operatorname{tg}\delta_2} \ll 1,$$

for this condition we have

$$C_x = \frac{R_2}{R_1} C_o. \tag{2}$$

In order to determine the deviation of the measured capacitance from the reference capacitance for a variable R_2 we have

$$\frac{C_x}{C_o} = \frac{R_2 - R_1}{R_1} 100\%. \tag{3}$$

We also obtain for

b) $$\frac{\operatorname{tg}\delta_x}{\omega C_x} = \frac{R_1}{\omega C_o R_2}\left(\operatorname{tg}\delta_o + \frac{1}{\operatorname{tg}\delta_2} + \frac{\operatorname{tg}\delta_o}{\operatorname{tg}\delta_2 \operatorname{tg}\delta_1} - \frac{1}{\operatorname{tg}\delta_1}\right).$$

By neglecting the term $\operatorname{tg}\delta_X / \operatorname{tg}\delta_2 \operatorname{tg}\delta_1$ whose value in our circuit is extremely small, we obtain

$$\frac{\operatorname{tg}\delta_x}{C_x} = \frac{R_1}{C_o R_2}\left(\operatorname{tg}\delta_o + \frac{1}{\operatorname{tg}\delta_2} - \frac{1}{\operatorname{tg}\delta_1}\right).$$

By inserting the value of C_X obtained from (2) we have

$$\operatorname{tg}\delta_x = \operatorname{tg}\delta_o + \frac{1}{\operatorname{tg}\delta_2} - \frac{1}{\operatorname{tg}\delta_1}.$$

Taking into consideration that $1/\operatorname{tg}\delta_1$ is a constant which is accounted for when the scale is calibrated, it is possible to omit it from the formula.

By denoting the loss angle readings on the scale by $1/\operatorname{tg}\delta_2 = \operatorname{tg}\delta_{SC}$ we obtain

$$\operatorname{tg}\delta_x = \operatorname{tg}\delta_o + \operatorname{tg}\delta_{AC} \tag{4}$$

If reference capacitor boxes with a loss angle < 0.001 are used with the DRCB instruments and capacitors with

a loss angle < 0.0001 with the PCB instruments, the value of $tg\delta_0$ can be neglected and it can be considered that $tg\delta_x = tg\delta_{sc}$.

The sensitivity of the circuit, which should remain constant in the unbalanced operating condition, will vary owing to variations in the bridge input impedance if capacitors of different nominal values are measured with a fixed supply voltage.

In order to keep the same tuning indicator percentage calibration scale for any nominal capacitance value, the sensitivity of the tuning indicator is adjusted for each nominal value. In mass-production measurements such an adjustment does not require much time, since large consignments of capacitors of the same nominal value are measured at a time. For such measurements the bridge supply voltage must remain constant.

For a convenient fixing of the measured capacitor the bridge can be supplied with a special contact device which will fit a capacitor of a given construction and will not provide additional losses.

The comparator bridges types DRCB and PCB(MKEP and MKT) have in practice provided satisfactory results with respect to accuracy and labor productivity, and should therefore find wide application in industry.

COMPUTATION OF DIFFERENTIAL RECTIFYING CIRCUITS

L. A. Sinitskii

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Differential circuits with rectifiers are used in devices for measuring the difference between two alternating voltages. These instruments can be divided into current and voltage measuring devices, and accordingly have two different circuits which we shall henceforth call series (Fig. 1a) and parallel (Fig. 1b) circuits.

The presence of a measuring instrument which has a resistance r_u upsets to a certain extent the operation of a differential circuit, since it provides a coupling between the two halves of the circuit which is contrary to the principle of operation of this device. One of the design problems consists of the correct selection of the instrument resistance for which the error due to the interaction of the two halves of the circuit is restricted to a preselected value.

Let us now examine from this point of view the series circuit. Let the compared emfs be equal respectively to

$$e_1 = E_{1m}\sin(\omega t - \varphi), \quad (1)$$

$$e_2 = E_{2m}\sin(\omega t + \varphi). \quad (2)$$

If the resistance of the measuring instrument were equal to zero, the current in circuit 1 would only depend on emf e_1, and in circuit 2 on emf e_2. The conducting time of each of the rectifiers in this case would be equal to half the period of the external emf, with rectifying 1 conducting during $\varphi \leq \omega t \leq \pi + \varphi$, and rectifier 2 conducting during $-\varphi \leq \omega t \leq \pi - \varphi$.

The mean value of the current in the measuring instrument

$$I_u = I_1 - I_2 = \frac{1}{r_r + r_f} \cdot \frac{E_{1m} - E_{2m}}{\pi}$$

(where r_f is the forward resistance of the rectifier) is determined by the difference in the amplitudes of the applied emfs and does not depend on the phase difference between them.

In the case when it is impossible to neglect the measuring instrument's resistance r_u, either of the currents i_1 and i_2 will be affected by both emfs (e_1 and e_2), and the currents i_1 and i_2 will be, during the time the two rectifiers are conducting, equal to:

$$i_1 = \frac{1}{r(2r_u + r)}[(r + r_u)e_1 + r_u e_2], \tag{3}$$

$$i_2 = \frac{1}{r(2r_u + r)}[(r + r_u)e_2 + r_u e_1]. \tag{4}$$

where $r = r_T + r_f$.

By introducing the notation:

$$E = \sqrt{E_{1m}E_{2m}}; \quad \gamma = \frac{E_{1m}}{E}; \quad k = \frac{r_u}{r + r_u}; \quad I = \frac{r + r_u}{r(r + 2r_u)}E$$

we shall obtain from (1) and (2) the following expression for currents i_1 and i_2:

$$i_1 = I\left[\gamma \sin(\omega t - \varphi) + \frac{k}{\gamma}\sin(\omega t + \varphi)\right], \tag{5}$$

$$i_2 = I\left[\frac{1}{\gamma}\sin(\omega t + \varphi) + k\gamma \sin(\omega t - \varphi)\right]. \tag{6}$$

After simple transformations, (5) and (6) can be represented in the following form:

$$i_1 = IA_1 \sin(\omega t - \varepsilon_1), \tag{7}$$

$$i_2 = IA_2 \sin(\omega t + \varepsilon_2), \tag{8}$$

where

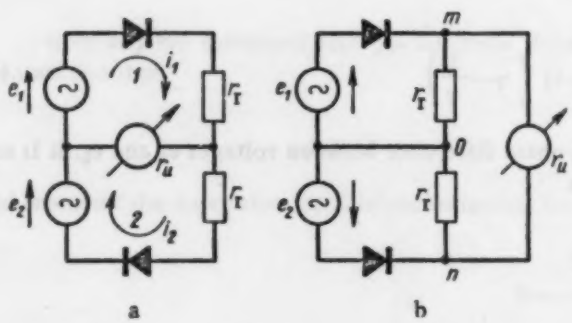
$$A_1^2 = \gamma^2 + \frac{k^2}{\gamma^2} + 2k\cos 2\varphi, \quad \operatorname{tg}\varepsilon_1 = \frac{\gamma^2 - k}{\gamma^2 + k}\operatorname{tg}\varphi,$$

$$A_2^2 = k^2\gamma^2 + \frac{1}{\gamma^2} + 2k\cos 2\varphi, \quad \operatorname{tg}\varepsilon_2 = \frac{1 - k\gamma^2}{1 + k\gamma^2}\operatorname{tg}\varphi. \tag{8a}$$

It is easy to see from (7) and (8) that rectifiers 1 and 2 will be conducting simultaneously in the time interval of

$$\varepsilon_1 < \omega t < \pi - \varepsilon_2,$$

since at $\omega t = \pi - \varepsilon_1$ rectifier 1 starts conducting, and at $\omega t = \pi - \varepsilon_2$ rectifier 2 stops conducting. Hence rectifiers 1 and 2 will be conducting separately in the time intervals of $\varepsilon_1 < \omega t < \pi + \varphi$ and $-\varphi < \omega t < \pi - \varepsilon_2$ respectively.



a b

Fig. 1.

According to the equalities which determine ε_1 and ε_2 these angles are smaller than φ. Hence each rectifier is conducting during a time interval greater than half the period of the external emf.

It is easy now to find the values of currents i_1 and i_2 taking into consideration the interaction between the two parts of the differential rectifying circuit. Let us note that during the time interval when only one rectifier is conducting, currents i_1 and i_2 are determined by the emfs e_1 and e_2 respectively, i. e.

$$i_1 = \frac{E_{1m}}{r + r_u}\sin(\omega t - \varphi) \quad \text{for} \quad \pi - \varepsilon_2 < \omega t < \pi + \varphi, \tag{9}$$

$$i_2 = \frac{E_{2m}}{r + r_u}\sin(\omega t + \varphi) \quad \text{for} \quad -\varphi < \omega t < \varepsilon_1. \tag{10}$$

By means of the previously adopted notations it is possible to express (9) and (10) in the form:

$$i_1=\gamma(1-k^2)I\sin(\omega t-\varphi),$$

$$i_2=\frac{1-k^2}{\gamma}I\sin(\omega t+\varphi).$$

The mean values of currents i_1 and i_2 are equal respectively to:

$$I_1=\frac{\omega I}{2\pi}\left[A_1\int_{\varepsilon_1}^{\pi-\varepsilon_2}\sin(\omega t-\varepsilon_1)dt+\gamma(1-k^2)\int_{\pi-\varepsilon_2}^{\pi+\varphi}\sin(\omega t-\varphi)dt\right]=$$

$$=\frac{I}{2\pi}\{A_1[1+\cos(\varepsilon_1+\varepsilon_2)]+\gamma(1-k^2)[1-\cos(\varphi+\varepsilon_2)]\};$$

$$I_2=\frac{\omega I}{2\pi}\left[\frac{1-k^2}{\gamma}\int_{-\varphi}^{\varepsilon_1}\sin(\omega t+\varphi)dt+A_2\int_{\varepsilon_1}^{\pi-\varepsilon_2}\sin(\omega t+\varepsilon_2)dt\right]=$$

$$=\frac{I}{2\pi}\left\{\frac{1-k^2}{\gamma}[1-\cos(\varepsilon_1+\varphi)]+A_2[1+\cos(\varepsilon_1+\varepsilon_2)]\right\}.$$

If we insert in the above equations the values of ε_1 and ε_2 obtained from (7a) and (8a), we shall obtain after certain transformations the following expressions:

$$I_1=\frac{I}{2\pi}[A_1+kA_2+\gamma(1-k^2)], \tag{11}$$

$$I_2=\frac{I}{2\pi}\left[A_2+kA_1+\frac{1-k^2}{\gamma}\right]. \tag{12}$$

The mean value of current in the measuring instrument circuit is equal to $I_u = I_1 - I_2$.

By inseting the values of I_1 and I_2 obtained from (11) and (12) we have:

$$I_u=\frac{I}{2\pi}(1-k)\left[(A_1-A_2)+(1+k)\left(\gamma-\frac{1}{\gamma}\right)\right]. \tag{13}$$

It will easily be seen that there is no interaction between the two parts of the circuit if the phase difference between the emfs e_1 and e_2 amounts to 180°, since in this condition each rectifier conducts during half a period, and the two half-periods do not overlap. For this case angle $\varphi = 90°$, $A_1 = \gamma - (k/\gamma)$, $A_2 = (1/\gamma) - k\gamma$, and the mean value of the current flowing through the measuring instrument is

$$I_0=\frac{I(1-k)}{2\pi}\,2(1+k)\left(\gamma-\frac{1}{\gamma}\right). \tag{14}$$

Since the measurement result should not depend on the phase difference between voltages e_1 and e_2, it is convenient to express the error of the above measuring device as

$$\delta=\frac{I_u-I_0}{I_0}.$$

From (13) and (14) the relative error is equal to

$$\delta=\frac{1}{2}\left[\frac{A_1-A_2}{(1+k)\left(\gamma-\frac{1}{\gamma}\right)}-1\right],$$

and, taking into consideration (7a) and (8a) we have

$$\delta=\frac{1}{2}\left[\frac{\sqrt{\gamma^2+\frac{k^2}{\gamma^2}+2k\cos 2\varphi}-\sqrt{k^2\gamma^2+\frac{1}{\gamma^2}+2k\cos 2\varphi}}{(1+k)\left(\gamma-\frac{1}{\gamma}\right)}-1\right]. \tag{15}$$

Normally the compared voltages differ little from each other, thus making parameter γ approach unity. Hence, it is of interest to find the limiting value of error δ when $\gamma \to 1$. By using l'Hospital's rule we obtain:

$$\delta_{\gamma=1}=\frac{1}{2}\left[\frac{1-k}{\sqrt{1+k^2+2k\cos 2\varphi}}-1\right]. \tag{16}$$

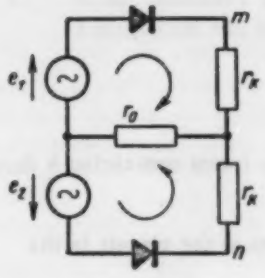


Fig. 2.

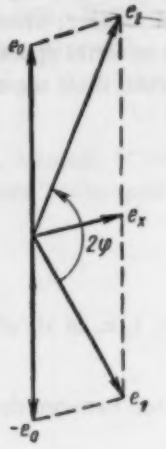


Fig. 3.

Normally r_u is relatively small, and hence, error δ is also small and (15) can be simplified. For this purpose let us represent the expressions under the radicals in the following form:

$$\gamma^2+\frac{k^2}{\gamma^2}+2k\cos 2\varphi=\left(\gamma-\frac{k}{\gamma}\right)^2+4k\cos^2\varphi,$$

$$k\gamma^2+\frac{1}{\gamma^2}+2k\cos 2\varphi=\left(\frac{1}{\gamma}-k\gamma\right)^2+4k\cos^2\varphi.$$

Owing to the smallness of r_u the value of $k\cos^2\varphi$ is small. Then on the basis of well-known approximate formulas we have:

$$\sqrt{\left(\gamma-\frac{k}{\gamma}\right)^2+4k\cos^2\varphi}\simeq\gamma-\frac{k}{\gamma}+\frac{2k}{\gamma-\frac{k}{\gamma}}\cos^2\varphi,$$

$$\sqrt{\left(\frac{1}{\gamma}-k\gamma\right)^2+4k\cos^2\varphi}\simeq\frac{1}{\gamma}-k\gamma+\frac{2k}{\frac{1}{\gamma}-k\gamma}\cos^2\varphi.$$

By inserting the above values of the radicals into (15) we obtain after simple transformations the following expression:

$$\delta\simeq\frac{k}{(1+k^2)-k\left(\gamma^2+\frac{1}{\gamma^2}\right)}\cos^2\varphi.$$

Normally the compared voltages differ but little from each other, i. e. parameter γ approaches unity. We can then assume that

$$\gamma^2+\frac{1}{\gamma^2}\simeq 2$$

and represent the expression for δ in the following form:

$$\delta\simeq-\frac{k}{(1-k)^2}\cos^2\varphi.$$

By neglecting k as compared to unity we finally obtain:

$$\delta\simeq-k\cos^2\varphi. \tag{17}$$

Let us now examine from the same point of view the circuit with a parallel connection of the measuring instrument (Fig. 1b).

For convenience of calculation let us transform the delta connection of two resistors r_r and resistor r_u into an equivalent star connection. As a result of this we shall obtain an equivalent circuit, shown in Fig. 2. According to the well-known conversion formulas resistors r_k and r_0 are determined by the following expressions:

$$r_k=\frac{r_u r_r}{r_u+2r_r},\quad r_0=\frac{r_r^2}{r_u+2r_r}.$$

The above equivalent circuit differs from the series circuit (Fig. 1a) by the direction of the rectifiers and the emfs. In the series circuit the emfs e_1 and e_2 are connected in series aiding, when the two rectifiers are conducting, whereas in the equivalent circuit of Fig. 2 they are then connected in series opposing. Hence, if equations (5) and (6) are used for determining currents i_1 and i_2 in the equivalent circuit, their second terms should be taken with a negative sign. Moreover, $\underline{k}$ in the circuit of Fig. 2 is equal to:

$$k=\frac{r_0}{r+r_0},$$

where $r=r_k+r_r$.

All the subsequent equations are changed in a similar manner, i. e. in all of them the terms containing $\underline{k}$ should have their signs reversed.

Thus, the formula for determining error δ due to the interaction between the two parts of the circuit in the case of a parallel circuit assumes the form:

$$\delta=\frac{1}{2}\left[\frac{\sqrt{\gamma^2+\frac{k^2}{\gamma^2}-2k\cos 2\varphi}-\sqrt{k^2\gamma^2+\frac{1}{\gamma^2}-2k\cos 2\varphi}}{(1-k)\left(\gamma-\frac{1}{\gamma}\right)}-1\right]. \tag{18}$$

For a small $\underline{k}$ and for γ approaching unity we have

$$\delta\simeq k\cos^2\varphi. \tag{19}$$

The above equations provide the values of the permissible error δ and of the parameter $\underline{k}$ for a given operating condition of the differential rectifying circuit (γ, φ). Thus, if the phase difference between emfs e_1 and e_2 can be given any value, cos φ may approach unity, and hence parameter $\underline{k}$ on the basis of (17) and (19) will satisfy the inequality

$$k<\delta_{max}, \tag{20}$$

where δ_{max} is the maximum permissible value of the error due to the interaction of the two parts of the circuit.

If the differential circuit is used in phase-sensitive divices [1], voltages e_1 and e_2 become respectively the sum and difference of the reference voltage e_0 and the measured voltage e_x (Fig. 3)

The phase angle φ in this case is determined with sufficient accuracy by the equality

$$\cos\varphi=\frac{E_{xrm}}{E_{om}},$$

where E_{xrm} is the amplitude of the measured voltage component at right-angles to the reference voltage; E_{0m} is the amplitude of the reference voltage.

Hence

$$\cos\varphi<\frac{E_{xm}}{E_{om}},$$

where E_{xm} is the amplitude of the measured voltage.

Normally $E_{xm} \ll E_{0m}$, i. e. φ approaches 90°.

By inserting the value of $\cos\varphi$ into (17) and (19) we obtain the condition for determining parameter $\underline{k}$ in phase-sensitive devices:

$$k < \delta_{max} \left(\frac{E_{om}}{E_{xm}} \right)^2 . \tag{21}$$

From (20) and (21) it is not difficult to find the maximum permissible resistance of the measuring instrument for which the relative error does not exceed δ_{max}.

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HIGH AND ULTRAHIGH FREQUENCY MEASUREMENTS

EQUIPMENT FOR CHECKING NONLINEAR DISTORTION METERS

B. K. Karavashkin

Translated from Izmeritel'naya Tekhnika, No. 7,
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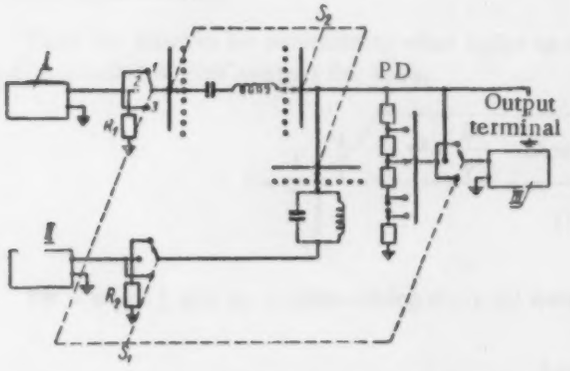
The Khar'kov State Institute of Measures and Measuring Instruments has developed an equipment for checking nonlinear distortion meters and spectral analyzers in the range of 60 cps to 15 kc.

Its principle of operation consists in supplying a voltage formed by two sinusoidal voltages of frequencies f and 2f. The nonlinear distortion coefficient in this case is equal to the ratio of the second and first harmonic voltages.

The equipment consists (see Figure) of two audio-oscillators I and II type ZG-12, a tube millivoltmeter III type MVL-2M, a filter unit and a stepped potential divider PD, in which the two oscillator voltages are added.



In position 1 of the three-pole switch S_1 the tube voltmeter III measures part of the oscillator I voltage taken off potential divider PD. Oscillator II is then replaced in the circuit by resistor R_2 whose value is equal to the oscillator's internal resistance. In position 2 of switch S_1 the voltage of oscillator II is measured by means of the tube voltmeter. The internal resistance of oscillator I is then replaced by an equivalent resistance R_1. If the oscillator I and II voltages are adjusted to obtain the same readings on voltmeter III, the ratio of the two voltages at the output terminals will be numerically equal to the ratio of the potential divider PD.

In position 3 of switch S_1 the voltages of oscillators I and II of frequencies f and 2f will be simultaneously fed to the output terminals, and the coefficient of nonlinear distortion of the total voltage will then be equal to the ratio of the potential divider PD.

Deviations from a sinusoidal form of the oscillator II voltage will have little effect on the nonlinear distortion coefficient error. With normally encountered nonlinear distortion coefficients of oscillators of the order of 0.5-0.7% this effect can be neglected.

The coefficient of nonlinear distortions of the frequency f voltage greatly affects the error in the output signal distortion coefficient. In order to reduce the above distortion coefficient the circuit includes m-type band filters.

The four-pole switch S_2 tunes the filters in its 9 positions to one of the following frequencies: 60, 100, 200, 400, 1000, 5000, 7500, 10,000 and 15,000 cps. These frequencies were selected for checking nonlinear distortion meter INI-10.

Nonlinear distortion meters with a continuously variable frequency within the audio range are also checked at these frequences owing to the above arrangement.

The filter bandwidth amounts to 3-5% of the resonant frequency. The filters provide an attenuation of the second harmonic by a factor of 7 and of the third harmonic by a factor of 10.

The filter inductors for frequencies above 1000 cps are air-cored. For frequencies below 400 cps the inductors are wound on toroidal ferrite cores (ferrite F-800, core cross section of 200 mm^2, length of the mean line-180 mm). All the coils are wound with PÉ-0.35 wire.

It has been established experimentally (by means of instrument ASChKh) that it is possible to neglect the distortions produced by the ferrite cores providing the voltage in the parallel circuit does not exceed 1 - 2 v.

The step potential divider resistors use manganin wire Ayrton-Perry windings on thin mica plates. The resistors are adjusted to their nominal dc values with a accuracy of ± 0.2%. The frequency error of such resistors does not exceed 0.2% in the audio range.

The tube voltmeter used in this equipment provides a 1% accuracy in comparing two voltages. The error in comparing two voltages of frequencies f and 2f is mainly due to the frequency error of the voltmeter and variations of its readings. Therefore, in selecting tube voltmeters for the above equipment special attention should be paid to these characteristics.

An analysis of this equipment's error components shows that the nonlinear distortions error of the output signal does not exceed ± 2% of the nominal value of ± 0.04% distortions.

In conclusion I should like to express my gratitude to L. D. Bryzzhev for guiding the work in producing the test equipment.

VARIATIONS WITH FREQUENCY OF STANDARD CAPACITORS

A. L. Grokhol'skii

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In metrological practice standard air capacitors certified only at 1000 cps are widely used as reference capacitances. Difficulties arise, therefore, when it is necessary to use reference capacitances whose effective values should be known at higher frequencies.

Owing to the absence of sufficient information on the behavior of such low-frequency standard capacitors at higher frequencies, the NGIMIP (Novosibirsk State Institute of Measures and Measuring Instruments) investigated the variation with frequency of the effective value of the "Étalon" plant type KVM and KVM-4 capacitors as well as of the capacitors made by Ulrich and Selinger. Fixed capacitors with values up to 1000 μμf were tested. Larger capacities were not tested, since they are very seldom used at higher frequencies.

Capacitor	Capacitance, μμf	Residual inductance, $h \times 10^{-9}$	Frequency in Mc at which the error does not exceed			
			0.1%	0.2%	0.3%	1.0%
KVM	50	9.01	7.5	11	13	24
	100	6.5	6.2	9	11	20
	200	19	2.6	4	5	8
	300	22	2	3	3.4	6
	400	7.5	3	4	5	10
	1000	14	1.3	2.0	2.3	4.0
KVM-4	100	32	2.8	4.0	4.9	8.9
	200	30	2.0	3.0	3.6	6.5
	300	36	1.5	2.2	2.7	4.8
	400	40	1.3	1.8	2.4	4.0
Selinger	200	7.5	4.0	6.0	7.0	13
	400	5.7	3.3	4.7	5.8	10
Ulrich	100	13	4.4	6.2	7.6	14
	200	16	2.7	4.0	4.8	8.8
	300	22	2.0	2.7	3.4	6.2
	400	21	1.7	2.5	3.0	5.5
	1000	28	1.0	1.3	1.6	3.0

The effective value of each tested capacitor was determined at several frequencies, depending on the nominal value of the capacitor. For instance, the highest frequency for the capacitor of 100 μμf was taken at 70 Mc, and for a capacitor of 1000 μμf at 10 Mc.

The processing of the results of these measurements leads to the conclusion that the equivalent circuit of an air capacitor can be represented up to 100 Mc by a network consisting of the low-frequency capacitance (the value allocated to the air capacitor in its 1000 cps certificate) and the residual inductance connected in series with it.

The value of the residual inductance for each of the tested capacitors remained constant within the errors of measurement.

The table shows the value of the residual inductances and the frequency errors of the tested air capacitors.

With the knowledge of the residual inductance it is possible to calculate the capacitor's effective value C_{eff}:

$$C_{eff} = \frac{C_n}{1 - l_k C_n \omega^2},$$

where C_n is the value allocated to the capacitor in its 1000 cps certificate; l_k is the value of the residual inductance.

The table also shows the highest frequencies at which the capacitor's effective value does not exceed its low-frequency value at 1000 cps by more than 0.1, 0.2, 0.3 and 1%. These frequencies were calculated by means of the above formula taking the residual inductance values obtained by measurement.

The values of residual inductances given in the table for the KVM and Selinger capacitors refer to their bottom terminating devices, and those for the KVM-4 and Ulrich capacitors to their top terminating devices, since owing to lay-out difficulties it was impossible to use their bottom terminals which are normally used at low frequencies. Attention should be paid to this peculiarity, since the use of either the top or bottom contacts in the capacitor determines the value of the residual inductances. Depending on the connection the inductances may differ from each other by 50 -70%. The residual inductances can vary between different capacitors of the same type and nominal value by about the same amount.

Keeping this remark in mind it is possible to assert that it is permissible to use for computing the effective values of capacitors of the same type the values of inductances given for those types in the table. This will obviously increase the errors in calculating the effective values, but the errors are not likely to exceed double the value they are given in the table, which is acceptable in the majority of cases.

In future, by introducing certain modifications in the manufacturing technique of our capacitors, it will be possible to obtain a reduced and constant value of the residual inductance, thus making it possible to raise accuracy in computing the effective value of the capacitors and to extend their frequency range. This will also provide the possibility of using in the application of capacitors nominal values of the residual inductances.

If higher accuracy is required than that indicated in the above investigations, it can be obtained by determining the effective value of each capacitor used for measurements. This is attained fairly simply by a complete substitution method, in which reference high-frequency capacitors are adjusted to be equal to the effective value of the capacitor under test. The possible errors in such comparisons can only be due to the instability of the measuring circuit and inaccuracy of its tuning, which can easily be kept within required limits.

For the above tests we used the high-frequency impedance bridge IS-1, but Q-meters and other suitable bridges can also be used for the above purposes. The reference high-frequency capacitor was made in the shape of a disc and could be adjusted to any value in the range of 25 to 1200 $\mu\mu$f [1].

The residual inductance of the tested capacitor can be found if required by processing the measurements of its effective capacitance and using the aforementioned formula.

LITERATURE CITED

1. A. L. Grokhol'skii, Izmeritel'naya Tekhnika No. 3 (1961).

ESSAYS AND REVIEWS

MODERN FOREIGN ANALYTICAL BALANCES

A. V. Marsov

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In recent years foreign firms have been producing most of their analytical balances with single pans, since these have many advantages as compared with double pans, and this has been duly recognized (see Table).

Some firms, for instance Sartorius (FGR) produce single-pan balances, but retain the beam unit of double-pan equal-arm balances with one main and two load-carrying knife-edge bearings, one of which is used to transmit the effort of the weights and the other to weigh the load. With such a design the variable error due to unequeal arms continues to affect the accuracy of balancing.

An essential and effective modification in the construction of anlytical balances has been introduced by the Mettler Company (Switzerland).

In the single-arm analytical balances manufactured by this firm the weighing is made at a constant maximum load with the weights placed on the same arm as the measured load, i. e., according to the principle suggested a long time ago by D. I. Mendeleev.

Weighing on one arm of the balances with a constant load stabilizes the scale graduations and completely eliminates the error due to unequal arms.

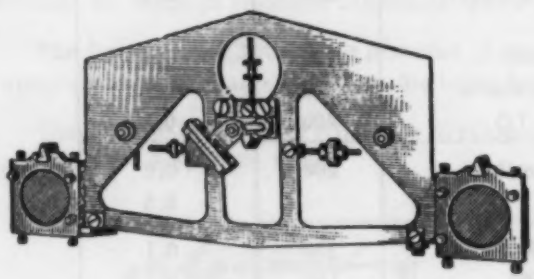

Fig. 1.

In their analytical balances H15 the Mettler Company have also improved the design of the weight-loading mechanism. The ring-type weights made of nonmagnetic chromium steel are placed horizontally in two rows. Such a loading mechanism provides a high reproducibility of readings, since the weights are placed on the pan without swinging.

The Ainsworth Company (USA) in their type TC analytical balances use vertically suspended weights which are deposited on the pan with precision and speed over the whole range up to 200 g.

With respect to the range of the optical scale the Mettler Company also occupies a leading position. In its analytical balances type VUS-1000 made in 1959 this scale amounts to 1150 mg.

Scales approaching this range are fitted to the Microva Company's (Switzerland) analytical balances, whose range amounts to 1010 mg and to the Christian Becker (USA) balances with a scale of 100 mg which is fitted with a micrometer vernier readout.

Such a readout on an optical scale speeds up the weighing and eliminates errors due to weights.

Other foreign firms are also extending their calibrated milligram scales, but their range is considerably smaller and the scale replaces only a minor part of the set of small weights.

Analytical balances type 707 of the nationalized plant of Oschatzer Wagenfabrik (GDR) have an interesting original design. These balances are not supplied with the usual pointer. The balance beam carries a mirror (Figs. 1 and 2). The deflection angle of the beam is transmitted by means of a light ray which is projected through an optical prism and a system of mirrors onto a scale on a vitograph screen. The length of the ray exceeds 0.5 m, i. e., it is considerably longer than the normal pointer, thus making even a very small deflection of the beam easily noticeable on the optical scale. A green filter can be placed in the path of the ray in order to reduce the strain on

Country	Firm	Single-pan analytical balances		
		Model	Maximum load, g	Calibrations of the beam deflections, mg
Britain	Stanton	Unimatic	200	0.1
	Towers	Ultramatic	200	0.1
		210	200	0.1
Hungary	Labor	4141	200	0.1
GDR	Oschatzer Wagenfabrik	707	200	0.1
	Herbert Witzke		200	0.1
	Franz Küstner	AR55	200	0.1
FGR	Zauter	Monopan	200	0.1
	Sartorius	Selekta-Standart	200	0.1
	"	Selekta-Rapid	200	0.1
	"	Proekta-Rapid	200	0.1
Holland	Becker Sons	Sitoptic	200	0.1
Poland	Mechanical Precision Plant	A-31	200	0.1
USA	American Balance Company	Quick Check	200	0.1
	Christian Becker	AN-1	200	0.1
	Voland	Unigramme	100	0.1
	Ainsworth	TC	200	0.1
France	Adamel	Vivansil	200	0.2
				0.5
Czechoslovakia	Meopta	A3-200	200	0.1
		A3-100	100	0.1
Switzerland	Mettler	B5	200	0.1
		H15	160	0.1
	Mikrova	Ultra-Rapid	200	0.1
		Exacta	100	0.2
Japan	Shimatzu	Libror	200	0.1
	"	"	100	0.1

the eyes. The replacement of a metallic pointer by a light ray greatly improves the display of the measurement results.

The latest analytical scales have a greatly improved weighing process. This is attained by effective damping, completely mechanized placing of weights, and an intermediate locking device. The time required for one weighing has been reduced to 3 to 6 seconds. For instance, in the Sitoptic analytical balances of the Becker Sons Company (Holland) the oscillations of the beam completely die out after 6 seconds, the beam oscillations of the Stanton (Britain) model A-50, A-48 and B-21 die out in 3-5 seconds, but the majority of the listed single-pan 200 g scale beams come to rest after 12-15 seconds.

The speed of the beam oscillation damping can be adjusted by varying the size of the throttle in the damping cylinder.

In the Release-o-Matic model of the Ertling Company (Britain) a pneumatic device is used which connects the isolating mechanism at a constant speed, thus ensuring a constantly smooth switching independently of the operator.

Certain firms apply the pan detainer which arrests its oscillations before isolating the beam. This preserves the zero position of the scale.

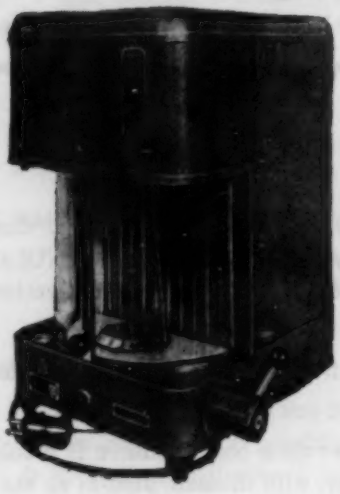

Fig. 2

The firm Adamel (France) produces original analytical balances with a maximum load of 200 g and a sensitivity of 0.2 mg (the Vivansil-357 model). This balance has no knife-edge bearings or blocks made of agate or sapphire, which are normally used in analytical balances.

The balance beam is made of invar and suspended from the balance column on 0.028 mm thick strips made of a special alloy. The cross pieces which carry the stirrups are also suspended on strips.

The analytical balances made by foreign firms often lack their characteristic column which rests on the base of the case. The column is replaced by a suspension which is fixed to the top of the case and serves as a bearing for the beam. With such an arrangement the weighing chamber becomes more spacious and convenient to use.

Special attention is paid for obvious reasons to the beam of the balance.

The use made by foreign firms of light nonmagnetic alloys for beams of analytical balances is a step in the right direction. Rolled nonmagnetic aluminum alloys are widely used. The firm Shimatzu (Japan) used for the first time in its Libror model balances made of titanium. In this balance the knife-edges and blocks are made of sapphires and rubies, whose temperature expansion coefficient approaches that of titanium.

The firm Towers (Britain) has provided its single-pan balances type 210 with a blocking device which prevents premature placing of gram-weights on the beam before it has stopped oscillating.

The firm Herbert Witzke (GDR) has developed a special device which prevents the slipping of blocks off the extreme knife-edges of the beam.

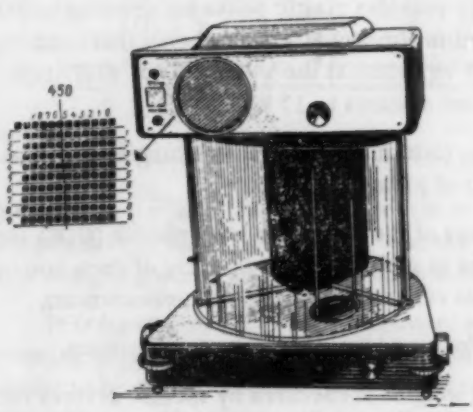

Fig. 3.

Special safety device are fitted to the beam knife-edges of analytical balances type Unimatic of the Ainsworth Company, type V-20 of the Stanton company, and others.

When the beam is locked it is also grounded to the frame by means of metallic contacts in order to reduce static charges. One pair of these contacts is rhodium plated.

In order to save time several firms provide preliminary tare balancing by means of pans with shot, special weights, etc.

The firm Oschatzer Wagenfabrik (GDR) provides for the suspension of pans a universal joint which ensures the horizontal position for the pan independently of the position of the weight on it.

The Mettler balances types H15 and H16 have hooks for suspending the load underneath the pans, thus providing the possibility of weighing, for instance, in a thermostat. The same firm makes single-pan analytical balances type B5 with a dismountable additional suspension for weighing under the balances.

The firm Franz Küstner (GDR) has produced in 1960 single-pan analytical balances type SAPSS with remote-control manipulation of the placing of weights, zero correction and locking. This balance is intended for weighing radioactive substances.

The analytical balance of the Christian Becker Company (USA) (Fig. 3) has a capacity of 200 g and provides gradual weighing, first with a accuracy up to 1 g, and then following this "rough" weighing it weighs with a accuracy of 0.1 mg with the reading taken off a vernier scale. The weight-placing mechanism has contacts with two signalling electric lamps inscribed "less" and "more."

The sighting line which is displaced along the vernier scale covers up the gap between two adjacent scale rectangles thus indicating tenths of a milligram.

In Fig. 3 the readout amounts to 455.4 mg. This graphical method of displaying the weighing results is very convenient.

Widescale standardization of the capacity of analytical balances, of their weighing cases and other components is characteristic of the present trend.

The knife-edge blocks are provided with an optical surface flatness finish. An indentation is made in the middle portion of the beam knife-edges.

Analytical balances are very sensitive to the effect of heat on badly lagged optical scale suspensions. Considerable improvements in thermal insulation of analytical scales has been attained by the Ainsworth Company (USA) in their SC model whose heat sources have been lagged and effective air-cooling provided by a simple constructional device.

Most of the weighing cases of foreign analytical balances have a metal frame and base, thus making them reliable in transportation and lowering their cost.

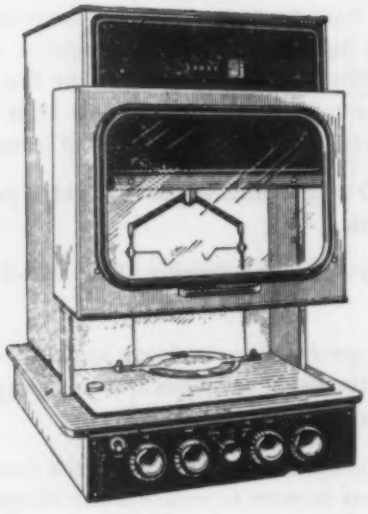

Fig. 4.

The latest analytical balances have their boxes finished in crackle enamel, epoxy paint or cellulose lacquer, with rhodium used as an anti-corrosive covering

The case of the Quick Check analytical balances (Fig. 4) of the American Balance Company is very convenient. It provides easy access to the weighing chamber since its front and both side walls are lifted up simultaneously.

Plastics are being used on an increasing scale. The firm Ainsworth (USA) makes in its Right-a-Way model plastic pans for the weight-placing mechanism, and plastic packing cases for transportation.

The Mettler company provides plastic boxes for covering up its analytical balance mechanisms (model H14 and others), thus reducing the weight of the balances to 8 kg, whereas the weight of our analytical balances of the "Gosmetr" plant amounts to 17 kg.

The Stanton Company (Britain) makes the weighing case of metal but the base of the balance of plastic material.

The 704-type balances of the Oschatzer Wagenfabrik (GDR) are provided with weights whose positive and negative errors are arranged in such a manner that any of their combined weights has an error approaching zero. Such an adjustment of weights raises the accuracy of measurements.

Foreign analytical balances are provided with flexible covers for an additional protection from dirt.

For transportation purposes the balance beam and incorporated weights are secured by special devices inside the case of the analytical balances. These packing components are painted in a special distinguishing colour.

In order to preserve the beam adjustment a special packing material, plastic foam, is used.

Packing details are made of multilayer corrugated cardboard and cut out to fit the balance weighing case. The entire cabinet containing the balance is also packed in this cardboard and then placed inside a packing case with a 10-centimeter layer of paper or wood shavings all around it.

The quality of laboratory lever balances is being raised not only in our country but also abroad, by means of new designs and a higher quality of workmanship.

Foreign firms pay considerable attention to the external finish of the articles, from the balances themselves right down to all the components and the packing. They aim to make every newly produced model of analytical balances of an original design.

The balances of the "Gosmetr" plant and especially those of the Khar'kov plant are inferior in external appearance to foreign balances, but it should be noted that in their metrological characteristics the "Gosmetr" plant balances are not inferior to those of many foreign firms.

The creative endeavors of Soviet designers and a further improvement in the production organization should raise the quality our analytical balances still further.

COMPLEX AC POTENTIOMETER AND ITS APPLICATION*

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The firm Hartmann und Braun (FGR) has developed a new rectangular-coordinate (complex) ac potentiometer by means of which it is possible to make most diverse measurements, including those of voltage, current, power, resistance, phase differences, ratios, etc. The potentiometer covers a wide range of frequencies, currents and voltages.

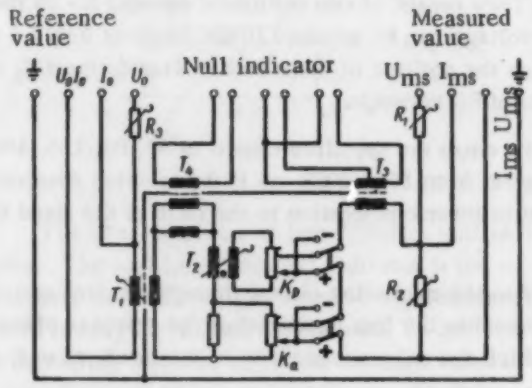

Fig. 1.

The potentiometer schematic is shown in Fig. 1. Measurements are made by the compensation method. The compensating voltage consists of the sum of two adjustable voltages with a phase difference of 90° between them, one of which is taken off the potential divider K_a which is supplied by the reference (operating) current I_0, and the other from potential divider K_p which is connected to the secondary winding of transformer T_2 whose primary winding carries current I_0. In order to be able to measure voltages in circuits with any potential to ground, the reference current circuit is provided with an isolating transformer T_1.

The unknown measured voltage U_{ms} is balanced out by manipulating the knobs of potential dividers K_a and K_p until the null indicator deflection becomes equal to zero. The U_{ms} component in phase with I_0 will then be equal to $U_{ms.a} = I_0 K_a$, and the component at right at right angles to I_0 will be equal to $U_{ms.p} = I_0 K_p$. The total voltage equal to $U_{ms} = I_0 \sqrt{K_a^2 + K_p^2}$ and the phase difference between U_{ms} and I_0 is found from the relationship $\tan \varphi = K_p / K_a$ where K_a and K_p are the readings of the respective potential dividers.

In order to provide this potentiometer with a wide voltage and current range it is equipped with a divider consisting of resistors R_1 and R_2 with an adjustable ratio. One of these resistors (R_2) can be used as an adjustable shunt. In order to be able to vary over a wide range the value of the reference current (or voltage) transformer T_1 is provided with a transformation ratio switch and fitted with an adjustable additional resistor R_3.

With this potentiometer it is possible to measure voltage, current and power by using the reference instrument (voltmeter or ammeter) which reads the reference voltage or current. The total error of such measurements is equal to the sum of the potentiometer and reference instrument errors. It can also measure the ratio of two quantities or the value of resistances by means of rough instruments for checking the reference voltage or current. The total error of such measurements depends only on the potentiometer error.

*Review of the article by G. Schlamp in "Der komplexe Wechselpstromkompensator und seine Anwendung," ETZ-A, 1960, No. 22

All these circuits are mounted in a common casing suitable for transportation. The nominal value of the reference current can be varied between 0.02 and 10 amp, with a power dissipation in the circuit not exceeding 0.6 va. The nominal reference voltage can be set in the range of 50 to 500 v with a circuit current not exceeding 0.02 amp.

Transformer T_1, which is fed by the reference voltage, has a toroidal core made of a magnetically soft material. Its windings are screened in order to prevent capacitative coupling. The secondary current of this transformer is fed to the inphase potential divider K_a and the primary winding of transformer T_2. The air-cored transformer T_2 is made in the shape of a toroid and enclosed in a screen of magnetically soft steel for protection against external magnetic fields. The transformer has tappings by means of which the secondary voltage is set to the same value for all the nominal frequencies.

TABLE

Type / Properties	Sensitivity, %/mm	Frequency range, cps	First harmonic filter	Voltage phase discrimination	Readings settling time, sec
Oscillographic null indicator	0.005	10-1500	Bandpass	Yes	0.1
Vibration galvanometer	0.05	15-400	Resonance	No	0.5
Tube voltmeter	0.2	10-2000	-	No	0.5
Cathode-ray oscilloscope	1.0	10-2000	-	Yes	0.1

Potential dividers K_a and K_p are similar to each other. They consist of two resistance decades 9 × 10 mv and 9 × 1 mv and a slide wire from 0 to 1 mv. The compensating voltage can be adjusted in the range of 0 to 0.1 v. The input resistance of the potential dividers does not depend on the position of their knobs. Transformers T_1 and T_2 have a constant load, their errors are compensated by means of RC networks.

The nominal fixed potentiometer frequencies for which its errors are specified consist of 50, 60, 150, 400, 800 and 1000 cps. The instruments can also be used at any frequencies from 10 to 2000 cps if the reactive component of the measured frequency is corrected after each measurement in inverse proportion to the ratio of the fixed frequency to that at which the measurements were made.

The voltages obtained from the potential dividers are fed to the measuring circuit through a switch which reverses their phases. The wwitch has four positions which correspond to the four quadrants of the complex plane. Thus an immediate indication is obtained of the quadrant in which the unknown quantity is measured, as well as its phase relation to the reference quantity.

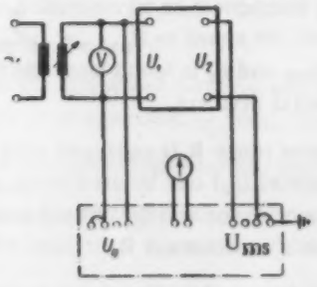

Fig. 2.

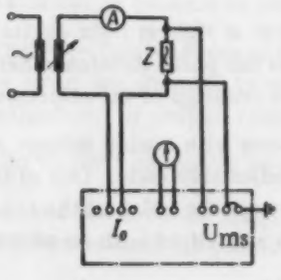

Fig. 3.

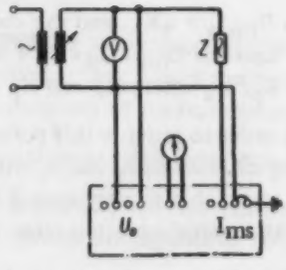

Fig. 4.

The resistive voltage divider is fed with the measured voltage in the range of 0.1 to 500 v, and the shunting circuit is calculated to take a current between 0.001 and 20 amp. The compensated voltage is then equal to 0.1 v. If the measured voltage is less than 0.1 v the voltage divider is not used. For a nominal measured voltage the circuit current amounts to 0.002 amp, and the voltage drop in that circuit at the nominal current is 0.12 v. The

measuring circuits can operate for a long time with voltages and currents exceeding the nominal by a factor of 1.2.

The measuring circuits are virtually unaffected by temperature, since their resistors are wound with manganin wire and their own temperature does not exceed 45°C. However, temperature variations do affect the reference circuits, since the copper windings of transformers T_1 and T_2 vary with temperature. These variations are compensated by means of thermistors which are in close thermal contact with the windings.

The nominal frequency switch has an additional position inscribed "demagnetization." This operation is required after electrolytic capacitor measurements involving direct currents which produce in the core of transformer T_1 residual magnetization which may result in additional potentiometer error amounting to 0.05%. The residual magnetization is eliminated by connecting transformer T_1 to a high load and feeding it with a current of 50 or 60 cps, decreasing in amplitude to zero.

The null indicator is connected to the potentiometer through a screened isolating transformer T_3. The null indicator can be either of a high- or low-resistance type with grounded or ungrounded input. For connection to the scanning plates of a cathode-ray oscilloscope transformer T_4 is provided.

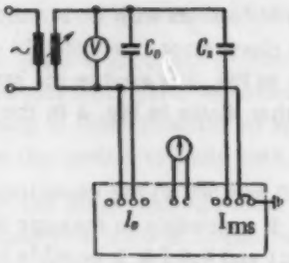

Fig. 5.

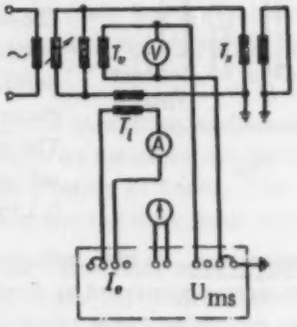

Fig. 6.

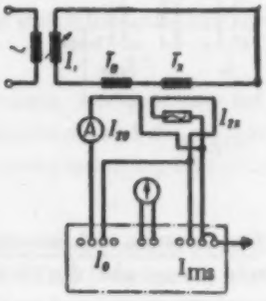

Fig. 7.

The characteristics of four different null indicators used with a complex ac potentiometer are shown in the Table. The oscillographic null indicator is the most sensitive of them. It also provides the possibility of balancing separately the resistive and reactive components at frequencies from 10 to 1500 cps. The vibration galvanometer is used mostly for frequencies between 15 and 400 cps, and owing to its high-frequency discrimination is convenient for measuring losses in steel cores at high value of induction. It is convenient to use the valve voltmeter and oscilloscope for frequencies above 150 cps. Their low sensitivity is not detrimental, since the potentiometer has larger tolerances in that frequency range.

The permissible potentiometer errors amount at 50 and 60 cps to ± 0.1%, at 150 cps to ± 0.2%, at 400 cps to ± 0.5%, at 800 cps to ± 1% and at 1000 cps to ± 1.5%. Phase errors expressed in radians amount to the same figures. When circuits are measured which are not affecred by the power consumed in the potentiometer input, the permissible errors are reduced to half the values given above.

For measurements at 50 and 60 cps the mains network is used for power supplies. In this case the normal variations in the mains voltage do not affect measurements, since both the reference and measured quantities vary in the same way, and their ratio remains constant. Frequency variations of the main voltage which can affect the accuracy in measuring the reactive component are also very small.

Measurements of voltage, current, power and angle. For accurate measurements of voltage (in the range of 0.05 to 500 v), currents (in the range of 0.0005 to 20 amp), and power (in the range of 0.05 va to 2 kva) a voltage stabilizer must be used, since the measuring instrument pointer will otherwise vibrate excessively. One of the potentiometer nominal standard voltages, for instance of 100 v, is chosen as a reference voltage or current and measured by means of a grade 0.1 instrument. The measurements are made as shown in the circuit of Fig. 2.

In measuring power, the mains voltage is measured with a standard voltmeter and serves as a reference value. The current is measured by the potentiometer. In checking voltmeters, ammeters and wattmeters the inphase po-

tential divider is first set to the correct value, and then the supply voltage is adjusted to obtain zero on the null indicator. In this case comparisons are obviously only made at the first harmonic.

Measurements of impedance and angle. The current flowing through the measured resistance is fed to one of the potentiometer inputs and serves as a reference current, the voltage drop across the impedance is picked up and measured on the other input. The complex ratio "voltage/current" is equal to the measured impedance. Its resistive component is read off the inphase potential divider, and the reactive component from the quadrature divider. The measuring circuit for this case is given in Fig. 3. The measuring range is 0.005 to 100 ohm. Thus it becomes possible, among other things, to measure contact resistances with large currents flowing through them.

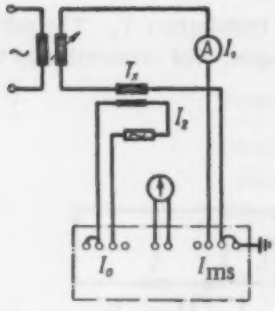

Fig. 8.

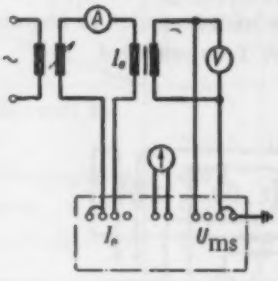

Fig. 9.

For measuring impedances between 25 ohm and 2 Meg the circuit shown in Fig. 4 is used. The power dissipated in the impedance can also be measured easily in this circuit. At 220 v it can amount to 0.2 to 2000 va.

Inductance measurements. It is possible by means of this potentiometer to measure the inductance of choke coils, coils, grounding circuits, as well as the mutual inductance of transformers with and without iron cores, including transformers with dc biasing. In these measurements the phase error is accurately determined. The circuit shown in Fig. 3 is used in the range of 0.15 mh to 0.3 h, and that shown in Fig. 4 in the range of 0.1 to 6000 h.

Capacitance and loss angle measurements. At high voltages (between 50 and 500 v) the capacitance and loss angle are measured with the potentiometer connected as shown in Fig. 4. It is possible to measure in this manner capacitances of 2 $\mu\mu$f to 100 μf. For measuring capacitances between 30 μf and 0.6 f it is possible to use the circuit shown in Fig. 3. Electrolytic capacities can be tested with the dc biasing voltage connected across them.

In addition, capacitances and loss angles can be measured by comparing the tested capacitor with a reference capacitor in a circuit connected according to Fig. 5. The advantage of this method consists in the simple calculations made in the course of measurements. By means of this circuit measurements can be made at high voltages, for instance, across insulators, windings of transformers and machines, etc. If the reference capacitance is taken at 100 $\mu\mu$f measurements can be made in the range of 2 to 50,000 $\mu\mu$f at a voltage of 100 kv.

Transfer ratio measurements. It is possible by means of this potentiometer to measure the input and output variables of any quadrupole and thus determine their ratio. For primary voltages of 50 to 500 v and currents of 1 to 10 amp it is possible to determine transfer ratios in the range of 1 to 2000.

Measurements of transformers and electrical machines. By placing on the core of the transformer under test an auxiliary coil with 1000 windings it is possible to determine the number of turns in the transformer coil in the range of 1 to 10,000. It is also possible to determine by means of this potentiometer the existence of any short-circuited turns in power transformer windings, open-circuit and short-circuit transformer losses, excessive temperature of windings, and insulation resistance. The circuit for measuring losses in a short-circuited power transformer is shown in Fig. 6. The circuit notations are: T_X is the transformer under test, T_U is the voltage transformer, and T_I is the current transformer.

Instrument transformer testing. By means of this potentiometer it is possible to check current and voltage instrument transformers. It is also possible to measure the transformer losses and its burden. Figure 7 shows the circuit for measuring instrument current transformers by a differential method, and Fig. 8 by a compensation method. The first method provides an accuracy of ± 0.001%, and the second of ± 0.1%.

Determining the characteristics of magnetic materials. By means of the circuit shown in Fig. 9 it is possible to determine the complex permeability of substances. The magnetizing current I_0 which flows through the winding of the sample also serves as a reference value for the potentiometer, the emf in the secondary winding of the sample constitutes the measured variable. Transformer steel samples can weigh between 0.1 and 50 kg. With samples weighing between 1 and 10 kg the measurement error amounts to ± 0.5%.

MATERIAL RECEIVED BY THE EDITORIAL BOARD

IMPROVING STATE INSPECTION OF MEASURING EQUIPMENT USED IN COMMERCE

V. V. Petropavlovskii

Translated from Izmeritel'naya Tekhnika, No. 7,
pp. 56-58, July, 1961

According to the August 9, 1960 resolution of the CPSU (Communist Party of the Soviet Union) central committee and the Council of Ministers, USSR on "Measures for a further improvement in trade," which noted considerable shortcomings in the operation of trading organizations, the Committee and its local agencies are faced with the task of improving state inspection of measuring equipment used in commerce.

The 1960 heavy inspection carried out by the State Inspection Laboratories (GKL) of the measuring equipment used in urban and rural trading revealed many measures and measuring instruments which had outdated state stamps, were out of order or provided inaccurate readings.

In order to radically improve the condition of measuring equipment in commerce it is necessary for the trading organizations themselves to enforce the legislation on measures and measuring instruments and the regulations on the supervision of these instruments approved by the Ministry of Trade. The instrument-repairing organizations should improve the quality of their work, reduce its cost and the time spent on repairs.

At the same time the local agencies of the Committee must organize efficient state inspection of the measuring equipment used in commerce. Correct organization of such inspection involves not only timely state testing and heavy inspection, but also constant and effective supervision by the GKLs over the administration of trading systems and instrument repair organizations.

Many GKLs are working in this manner with the trading organizations, with the object of improving and exextending the activity of service inspection agencies, by summarizing the data obtained in heavy inspections, organizing conferences, lectures and engaging in other similar activities. However, despite all these measures the condition of the measuring equipment in commerce as a whole is not improving. This is due to the fact that the organizational work is far from being carried out by all the GKLs, and in the majority of cases it is done unsystematically, without a definite objective, and without initial data on the existence and condition of measuring equipment in various trading departments.

Many heads of GKLs underestimate the importance of such inspection work, forgetting, in particular, that the number of inspection items in urban and rural trading amounts to some 50% or more of the total number under their control, and hence, the condition of the commercial measuring equipment has a decisive effect on the state of measuring equipment in general in a given region, territory or republic.

It should also be stressed that the organization and carrying out of state inspection of measuring equipment in rural districts often presents many difficulties, involving travelling, weather difficulties, transport condition, travelling expenses, difficulties in storing and using the equipment, less qualified personnel in the commercial establishments, etc. This requires considerably greater efforts and expenditure than inspection of urban trading.

It will be seen, however, from the 1960 reports that in several administrations of the Committee, for instance, in the Latvian, Lithuanian, RSFSR, Uzbek, Estonian and other areas, the percentage of defective instruments in urban trading is higher than that of the Tsentrosoyuz (Central Union of Consumers' Cooperatives) which operate in rural areas. This provides grounds for the assertion that the heads of many GKLs as well as the Committee's representatives in some of the Union republics pay little attention to the efficient organization of the state inspection of measuring equipment in the urban trading organizations of their republics.

An important organizational defect of this inspection consists in the insufficiently vigorous heavy inspection, which indicates not only a lack of discipline in the trading organizations, but also low efficiency of many GKLs in checking the implementation of previously made suggestions.

In order to organize more efficient and effective state supervision of the measures and measuring equipment used in commerce the following measures should be undertaken. First of all a specific plan for the entire organizational and inspection work should be compiled, indicating the objectives, means and times of its application. The plan must differentiate strictly between heavy inspection of urban and rural trading, since the methods and time limits involved in either case are totally different.

The first heavy inspection of urban trading should be carried out as an administrative inspection in the first quarter of the year. It should be prepared by organizing beforehand the required conferences, training, snap inspections in the presence of trading organization representatives, and other measures.

On completing the inspection of urban trading the data should be duly classified for each department. The data on defective instruments should be carefully analyzed. If, for instance, the majority of defective instruments have outdated state stamps, it is an indication of unsatisfactory work of the GKLs in the previous year. If, however, the majority of the defective instruments have been classified as such, owing to their dirty and faulty condition, this points to the inefficiency of service inspection and unsatisfactory training of commercial workers in the maintenance of the measuring equipment, etc. Experience has shown that shop managers can, with proper instruction, become useful assistants to the GKLs in the efficient maintenance of measuring equipment. For instance, the manager of provision shop No. 30 of the Tambov Gorpishchetorg (City food trading organization), I. M. Sychev, carefully inspected each morning the state of all his weighing equipment (correct positioning, cleanliness, state of preservation, efficient operation, etc.). Each shop assistant is made strictly responsible for the state of the scales and weights entrusted to him. Therefore, the weighing equipment in this shop is always in a good condition and there are no complaints from customers about incorrect measuring or weighing out of goods. The expenditure on the repairs of scales and other instruments has been considerably reduced. On com. Sychev's request the instruction of the workers in shop No. 30 on the handling of weighing equipment is made directly in the shop, thus making it easier for them to assimilate the minimum technical knowledge required. As yet, however, there are only a few such shop managers, and the GKLs are exerting insufficient efforts in this respect.

After working out the inspection results it is necessary to send appropriate reports to the organizations concerned, and if necessary to the regional or city trade administrations and other city organizations, insisting on the fulfilment of the proposed recommendations. Sometimes it is useful to write to the local press on the inspection results, or to make a report at the meeting of the appropriate city Soviet executive committee, insisting on the adoption of the required measures and on their fulfilment.

As far as state inspection of measuring equipment in rural trading is concerned, it requires even more careful preparation, both with respect to primary and heavy inspections.

In many GKLs the above inspection is carried out in the following manner: temporary branches as the main method of this inspection are organized every other year, and heavy inspection of half the equipment every year. Often the specific inspection plan falls through and the head of a temporary branch only inspects those establishment which are near the district center. As a result of this, very often the same establishments which are near the district center are inspected from year to year, whereas those further away from the center are not inspected for several years.

It is necessary to have for every rural district a complete list and location in every inhabited locality of the establishments equipped with measuring equipment. The existence of such lists will make it possible for the head of a temporary branch to plan in advance the itinerary of his test and inspection visits in a given district. The work in the districts will become more productive and better organized. It will become possible to cover every two years all the establishments of the district, thus providing accurate data on the condition of the establishments and their measuring equipment. Then it will become possible to summarize the inspection data in rural districts in the same manner as in urban didstricts, including the data on rural trading, and to take the necessary measures by reporting to the regional (territorial) organizations, etc.

If side by side with the above measures the GKLs will exert the required pressure on the administration in charges of the technical and service inspection of measuring equipment, and improve the work of instrument repair shops, the ground will be prepared for a rapid and considerable improvement in the condition of measuring equipment used in commerce.

The state inspection laboratories must in every way assist the extended operation of the service inspection agencies, bearing in mind that the main indication of their activity consists in the number of instruments they

inspect. At the same time it is necessary to demand high-quality repairs of weighing equipment by the instrument-repair organizations, paying attention to such things as external appearance of scales (high-quality painting, nickel plating, etc.), especially for instruments used in trading. It is often the case that newly repaired dial scales which are delivered to a shop have serious defects and cannot be normally used. This shows not only the low quality of their repairs but also the inefficiency of state inspection of repaired instruments.

It is necessary to organize an exchange fund of instruments, which is an efficient method of organizing repairs. This method can be applied in two ways, a passive and an active manner. The passive manner involves exchange of defective instruments for repaired ones which have passed their state test and have been stamped at the instrument-repair organization. In the active manner the instrument-repair organization itself (normally a service repair organization) carries out this exchange is very convenient both for the trading organizations and the GKLs and will help to improve the condition of measuring equipment in commerce.

Attention should also be drawn to the clear stamping of scales and especially weights. Unclear stamping leads to many difficulties, both for the trading organizations and the GKLs, raising the number of defecting instruments in circulation.

The above does not cover all the organizational measures which could be taken by the GKLs for the improvement of state inspection. On the basis of the condition and organization of measuring equipment in each region and territory, the laboratories must work out concrete measures for the improvement of state inspection in trading, and insist on a decisive improvement of the measurement technique in their region. It is advisable to involve social organizations in this work (such as the Komsomol crews, workers' inspection in trading organizations, assistance groups of pensioners, etc.). This work will be rendered much easier for the GKLs of the Russian Federation in view of the recent instruction of the RSFSR Ministry of Trade establishing the "Regulations on the service inspection of measures and measuring instruments in the organization of the Ministry of Trade, RSFSR," and the "Rules for using measures and measuring instruments in trading and public catering establishments."

A radical improvement of state inspection of measuring equipment in trading is for the GKLs a task of state importance which must be fulfilled in the near future, as required by the resolution of the CPSU central committee and the Council of Ministers, USSR on the "Measures for a further improvement in trading."

TRAINING ELECTRICAL ENGINEERING STUDENTS IN ELECTRICAL MEASUREMENT TECHNIQUES

A. M. Turichin

Translated from Izmeritel'naya Tekhnika, No. 7,
pp. 58-60, July, 1961

The training of electrical engineering students in electrical measurement techniques is provided under the present plan of higher educational establishments by a general course in electrical measurements. The assimilation of the required knowledge in electrical measurement techniques is very important in forming the scientific and technical outlook of students who are preparing to become electrical engineers.

Any study course must provide the student with a sum of knowledge on the level of modern scientific and technical development in this sphere, as required by the student for this future activity as an engineer.

The present curriculum and available textbooks of the general course on electrical measurements do not satisfy these requirements.

Let us examine this question with respect to the three basic electrotechnical specialities: a) power stations; b) electrical machines, and c) electrical equipment of industrial establishments.*

*This article does not refer to students specializing in "electrical measurement technology" or "automation and remote control." for whom the courses on electrical measurements are only formative.

An electrical engineer working at a thermal power station (and especially at an atomic station) deals not only with instruments measuring electrical quantities (current, voltage, power, etc.), but, in fact, he deals mostly with instruments measuring and controlling flow, level, temperature and the concentration of liquids and gases (control of the flue gas composition, temperature, bearing vibrations, salt content in condensed water, etc.) and many other nonelectrical parameters which characterize various stages in the production of electrical energy.

An electrical engineer working in an electrical engineering plants must be familiar not only with electrical measuring equipment used for checking the electrical parameters of machines, but also with the equipment used for checking vibrations and vibration resonance, for static and dynamic balancing of rotors, for measuring axial displacements and efforts, for measuring mechanical tensions (strain gauges), for measuring temperature, for checking hydrogen content in hydrogen cooling systems, and for checking many other nonelectrical parameters of electrical machines.

An electrical engineer specializing in the electrical equipment of industrial plants must be familiar with instruments for measuring mechanical parameters of lathes and geometrical parameters of the produced commodities, i. e. of all the commodities which can be used as a basis for the automation of electrical drives.

From the above far from complete enumeration of the problems in measurement techniques forming the sum of the knowledge which should be communicated to the student in a general course on electrical measurements it becomes perfectly obvious that the curriculum should include measurements of electrical and nonelectrical physical quantities.

If the course on electrical measurements in considered as a system embracing the theory and practice of electrical measurement technology and its application in industry and science, it then becomes necessary to include in it measurements by electrical means of all physical quantities.

The existing syllabuses and textbooks of a general course in electrical measurements deal in the main with electrical measurements of electrical and magnetic quantities. The section on electrical measurements of nonelectrical quantities (temperature, geometrical and mechanical quantities, liquid and gas concentrations) is relegated to the end of the syllabuses and textbooks and given very little space. Lecturers, as a rule, allocate to this subject not more than the last four hours of their lectures. The syllabuses on laboratory experiments in the course, in the majority of institutions for higher technical education, do not contain any practical work on measurements of nonelectrical quantities.

It is quite obvious that a mechanical extension of the number of lecture hours and the volume of textbooks dealing with the method and describing the apparatus used in measuring all the physical quantities is, in practice, impossible. This problem can only be solved by a qualitatively new approach to the exposition of this very extensive material.

Modern electrical measuring instruments cover a very wide range and are often very complex in design. However, no matter how complex the instrument may be it can always be represented as a combination of separate transducers which convert one quantity into another. The input of a chain of such transducers is fed with the measured physical quantity, and its output is connected either to a readout device or an actuating mechanism which operates without human participation.

No matter how diverse the combinations of transducers forming various electrical measuring instruments may be, the variety of transducers distinguished by their types, i.e., by their principle of operation, is restricted.

Thus, it is perfectly natural to divide a general course on electrical measurements into two parts: a) transducers as elements of any electrical measuring device, and b) measurements of various physical quantities: current and voltage, electrical power and energy, parameters of electrical circuits (R, L, M, C, f, φ), temperature, geometrical and mechanical quantities, liquid and gas concentrations, and magnetic quantities.

In the first part of the course it is advisable to divide the transducers into the types of energy conversions: a) transducers of electrical quantities into the rotation of a moving part (measuring mechanisms); b) converters of electrical quantities into other electrical quantities (current and potential dividers; instrument, matching and isolating transformers, ac to dc converters and vice versa, electronic amplifiers, etc.); c) transducers of nonelectrical into electrical quantities; d) transducers of magnetic into electrical quantities.

In the second part of the course (measurements of various physical quantities) it is only necessary to deal with

the principal methods and instruments used in measuring any given quantity, and to describe them as an example of the combination and utilization of the transducers which have been studied in the first part of the course.

Such a presentation of the material makes superfluous the separation of the course on electrical measurements into two parts, namely one dealing with measurements of electrical, and the other with nonelectrical quantities. This separation arose historically and is now obsolete.

In dealing with this vast material both at lectures and in textbooks it is necessary to sacrifice the design peculiarities of instruments (for students of the above mentioned specialities who are not training to become instrument designers) in order to reduce the size of the course (and textbooks). The number of lecture hours dealing with a general course in electrical measurements according to the above syllabus should be not less than 60, and the number of hours spent on laboratory experiments not less than 48. The fact that in syllabuses for students of various electrotechnical specialities the number of lecture hours dealing with general electrical measurements has been reduced to 32-36, and the number of hours spent in laboratory experiments to 30, deals an serious blow to quality in training electrical engineers fit for the present and especially the future state of our industry.

A textbook on a general course of electrical measurements should not exceed 20 quire (1 quire per 3-4 lecture hours), so as to enable students to prepare for the examinations from the book instead of using notes or lecture abstracts. This is particularly important for students in evening or correspondence classes.

In conclusion we provide a tentative distribution of lecture hours for various subjects in the course.

Introduction		1
1.	Transducers of electrical quantities into the rotation of a moving part (measuring mechanisms)	6
2.	Converters of electrical into other electrical quantities	8
3.	Transducers of nonelectrical into electrical quantities	8
4.	Transducers of magnetic into electrical quentities	3
5.	Current and voltage measurements	5
6.	Measurements of electrical circuit parameters	8
7.	Electrical power and energy measurements	4
8.	Temperature measurements	4
9.	Measurements of geometrical and mechanical quantities	4
10.	Measurements of liquid and gas concentrations	6
11.	Measurements of magnetic quantities and ferromagnetic material testing	3
	Total	60

ANNOYING DEFECTS

S. M. Rovinskii

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pp. 60, July, 1961

The Vinnitsa butter and fat combine is using instruments made by the Moscow "Manometr" and "Énergopribor" plants and a L'vov plant. These instruments have defects which at first sight appear to be trivial, but make the use of the instruments more difficult and involve unproductive expenditure of time.

Thus, the Moscow "Manometr" plant makes a differential transformer instrument type ÉPID and transducers type DM-6. The set of spares delivered with these instruments includes electron tubes, pens, inkwells, stranded wires, and even pipettes. None of these components is in short supply and can be obtained relatively easily. The most important component of the set consists of the differential manometer DM-6 diaphragm unit. A defect in this unit puts the entire flowmeter out of action. It appears from our correspondence with the "Manometr" plant that

it does not supply this spare part. Diaphragm units fail to operate fairly often. Thus, in 2 of the 13 instruments received by us and put into operation the diaphragm units have failed. It is very essential to include a diaphragm unit in the set of spare parts.

The nonmagnetic steel tube which holds the plunger of transducer DM-6 is secured by a special lock nut. In adjusting the instruments we have had to dismount some of the tubes in order to ascertain the condition of the diaphragm unit. The lock nut was tightened up hard but made of a relatively soft material. Repeated attempts to unscrew the nut by means of end and box spanners with tommy-bars rounded off the sides of the nut, so that eventually a cold chisel and hammer had to be used to remove it. A similar effect was obtained with the drainage pipes in the positive and negative chambers of differential manometer type DM-6. It is necessary to change the material of the nuts and drainage pipes and enlarge the width of the faces which take the spanner.

Experience in the use of automatic electronic bridges MS1-01 of the L'vov plant shows that the inkwells with their platino-rhodium pens are virtually useless. The combine in the last 2 years has been using 20 such bridges. Repeated attempts to make these pens work have not produced any results, they either refuse to write or else the ink comes pouring out in a stream. Only 2-3 worked relatively satisfactorily. Finally we replaced the pens with short pencils, fixing them in the place of the inkwells. Is it really necessary to make all the recordings in ink? The use of slate pencils would considerably simplify the design and maintenance of these instruments and provide a saving of expensive metal.

If, however, it becomes necessary to record in ink, perhaps a tube with printers' ink and a ball tip similar to a ball pen could be used.

The supply lead of the instrument MS1-01 and the lead from the resistance thermometer are fixed inside the set; due to the considerable depth of the instrument this is very inconvenient, since the connection has to be made blindly by fumbling with one hand. A terminal box with a detachable lid should be provided at the rear of the instrument. During the use and repairs of the instrument its electrical circuits have to be checked. The plant supplies instructions and wiring and assembly drawings, on which all the wires and components are numbered, but it is impossible to follow these drawings. With the 20 instruments at our disposal we have been unable by comparing them to follow this numbering. The wiring diagram should be exceptionally clear and legible.

The VTI electronic regulators of the Moscow "Énergopribor" plant have a relatively complicated circuit, but no assembly and wiring instructions or diagrams are supplied, and the components and wires (with the exception of the external terminal blocks) are not numbered.

The elimination of the above-mentioned defects will make the use of these instruments considerably easier.

SUPPLYING IN TIME THE REQUIRED OFFICIAL LITERATURE TO THE GKLs (STATE INSPECTION LABORATORIES)

N. A. Pleva

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pp. 61, July, 1961

The experience in supplying official documents and specifications of the Committee of Standards, Measures and Measuring Instruments to the GKLs shows that newly approved instructions, regulations and specifications on measuring equipment, as well as their supplements and amendments, are implemented with considerable delay, owing to their belated appearance, and some of these never even reach state and service inspectors.

Besides their belated receipt, official documentation is often too numerous and uncoordinated; furthermore, there does not seem to be any thought-out plan for its publication and bringing it to the information of inspectors. It all boils down to notifications of the approval and publication of such literature, instead of its practical utilization within the conditions of the GKLs.

The East Kazakhstan GKL employs 16 operative workers. The Leninogorsk and Zyryanovsk area permanent branches, which have 7 operative workers, form part of the GKL. But under the Committee's allocation we only get 3 copies of any regulations and instructions that are published, and only 1 copy each of alterations and amendments to instructions.

As for the numerous workers in the service inspection agencies, they are completely deprived of the possibility of acquainting themselves with additions and amendments that are being put into effect until the appropriate official documents are republished and the amendments incorporated in them.

In all probability the officials of the Committee consider that the GKLs should themselves duplicate and send out all official documents of the Committee to all inspectors. However, this is an impossible task. Moreover, it is fundamentally incorrect, for errors may occur in the reprinting of documents.

There is a practice in force at present whereby the state inspector must bear the cost of subscriptions to official publications for service use. This practice, however, can only be obligatory in respect of technical literature, and not in respect of the official publications of the Committee, and is totally inapplicable to amendments and changes to instructions and regulations. Moreover long periods, sometimes streetching into years, elapse between the publication of official literature and its practical application. All this time the inspectors have to work according to obsolete documents which in actual fact have been revoked.

For instance, regulation No. 5 of January 20, 1958 provides amendments and supplements to instructions Nos. 40-56, 57-56, 66-56 and 69-56. Since this regulation occupied 16 printed pages we were unable to produce it in 9 copies required by our workers who are employed in checking balances and weights. Only by the middle of 1960 did we receive the required quantity of new instructions with the amendments incorporated in them.

We consider that it would be better and more efficient for the Committee to dispatch to the GKLs all the official documents not in a haphazard manner but according to the number of inspectors employed in the given GKL.

Additions and amendments to regulations should be issued in the form of loose leafs and dispatched in quantities sufficient to cover not only state inspectors but also service inspection workers. Moreover, it is not necessary to quote the resolution which introduces an amendment, it is sufficient to refer to its number and date of issue.

Finally, the Committee should supply its informative and official literature from a central source to the workers employed in area measuring equipment branches.

One should also like to raise the question of issuing compact measuring equipment handbooks.

Editorial note. The question raised by the author is of considerable importance. The readers of this journal expect the Measuring Equipment Administration to inform them on the steps taken in this connection.

INFORMATION

SESSION OF THE GENERAL ASSEMBLY AND THE COUNCIL OF THE INTERNATIONAL STANDARDIZATION ORGANIZATION

Translated from Izmeritel'naya Tekhnika, No. 7,
pp. 62, July, 1961

In June 5-16, 1961 sessions of the General Assembly and Council of the International Standardization Organization (ISO) as well as of its 10 technical committees and subcommittees was held in Helsinki (Finland). Delegations from 38 member countries attended the ISO session in Finland. The total number of participants amounted to some 600 persons, including (according to the secretariat data) delegations of 72 persons from Great Britain, 79 from Germany (GDR and FGR), 66 from France, 54 from Sweden, 39 from USA, 28 from the USSR, 25 from Italy, and 25 from the Netherlands.

According to the ISO statutes the General Assembly consists of representatives of the member countries' national organizations engaged in standardization. It is called every three years to deliberate on the most important questions of its activity. At present the ISO consists of 44 countries including all the European socialist countries.

In the intervals between the sessions of the General Assembly the Organization is controlled by a Council, consisting of a president and representatives of 14 member countries, which is elected every three years. In order to preserve the continuity of operation the term of office of the council members does not end simultaneously, so that every year not more than 4-5 members are reelected.

At present member countries of the ISO Council consist of Austria, Great Britain, Germany (Germany is represented in the ISO by a united delagation of the GDR and FGR standardization organizations), Israel, Italy, the Netherlands, Norway, Poland, Rumania, The USSR, USA, France, Switzerland and Yugoslavia.

The Council elects a vice-president, treasurer and a third member of the control committee who, as a member of the Council assists the vice-president and the treasurer in supervising the activity of the General Secretariat.

In order to help the Council in its many-sided activities several general committees, such as planning, editorial, finances, control and other committees are established.

The ISO Council at its session dealt with the following basic problems: elections of the vice-president, treasurer and the third member of the control committee; appointment of a financial inspector for 1961; consideration of the General Secretariat reports on the work of the technical committees on implementing the ISO recommendations for national standards; relations with other international organizations, etc.

J. Birlet, the director of the French Standardization Association, who was nominated by the Soviet delegation, was elected vice-president of the ISO. J. Wodzicki, president of the Polish Standards Committee, was elected the third member of the control committee. For the second time running de Sagy (Switzerland) was elected treasurer. The Swiss Financial Control Society, which for several years has audited the ISO finances, was appointed financial controller.

The ISO Council approved the reports of the General Secretariat, which stated that at present over 100 technical committees are operating in the ISO, 200 ISO recommendations have been approved, and over 400 draft recommendations have been submitted for approval. Relations with many other international organizations have been established. The ISO recommendations are beginning to be adopted in the national standards of many countries.

The leader of the Soviet delegation, A. E. Vyatkin, noted the improved work of the General Secretariat and reported that over 100 ISO recommendations have been either fully or partially adopted in State Standards of the Soviet Union, a large number of recommendations will be incorporated in Soviet standards in the near future, and only the recommendations which specify lower requirements than those provided by Soviet standards will not be adopted in the GOSTs (All-Union State Standards).

The Council approved several ISO recommendations, agreed to the formation of four new technical committees, and adopted a resolution on the improvement of ISO activity.

The General Assembly opened on June 15, 1961. The President of Finland, U. Kekkonnen, members of the Diplomatic Corps, including the Soviet Ambassador to Finland A. V. Zakharov, were present at the opening.

The main items on the General Assembly agenda consisted of the election of the ISO President and of the five new members of the Council.

A. E. Vyatkin, chairman of the Committee of Standards, Measures and Measuring Instruments attached to the Council of Ministers, USSR was unanimously elected President of the ISO for 1962-1964. In a secret ballot the Soviet Union, India, Czechoslovakia, Germany (GDR and FGR) and New Zealand were elected members of the ISO Council.

In conclusion it should be noted that at all the sessions of the ISO in Helsinki the work was conducted in atmosphere of friendship, understanding and respect.

K. D. N.

FROM THE JOURNALS

JOURNAL OF SCIENTIFIC INSTRUMENTS

No. 5, 1961

A. M. Taylor. The annual exhibition of the Institute of Physics and the Physical Society, London, 1961.

Description of the instruments shown at the exhibition, including: an interferometer of a special design, a photoelectric spectropolarimeter, radiomicrometer, high-temperature thermocouples, etc.

J. A. Hall and V. M Leaver. Stabilization of thermometers of borosilicate glass for use at high temperatures.

Investigation of the best heat-treatment method for borosolicate glass thermometers before their calibration, with the object of providing high stability in their use in the range of 400 to 500°C.

B. Yates. Indium resistance thermometers.

Description of the thermometer design which consists of an indium wire and is especially suitable for cryostats with a confined operating space.

J. Middlehurst. Photoelectric optical pyrometer.

Description of the optical pyrometer in which brightness is controlled by a photoelectric method. Its sensitivity is ± 0.02°C at a temperature of 1063°C, and its reproducibility is ± 0.1°C.

FEINGERÄTE TECHNIK

No. 4, 1961

Establishment of a German society for measurement techniques and automation.

A society has been established in the GDR with the object of developing in every possible way measurement techniques, automation and instrument making, and strengthening international cooperation in these spheres.

Conference on problems of quality, organized by the administration of the national establishment "Control Techniques, Instrument-Making and Optics," was held in Leipzig in January, 1961.

Drescher. Measuring and control instruments at the Leipzig Spring Fair, 1961.

Tromler. Sonivisor 2. Description of the design and principles of operation of an ultrasonic instrument for testing materials.

H. Weinhold. Measurement of cylindrical gears by means of balls, part 2. Practical examples.

No. 5, 1961

Blume. Group processing of components by Mitrofanov' method and its application at the national plant of Karl Zeiss, Jena.

Hertel. The present state of surface quality standardization.

R. Schneider. Production considerations in introducing ISO (International Standardization Organization) metric threads.

Haas and Padelt. Comments on the decisions of the 11th General Conference on Weights and Measures (Paris, October, 1960) on determining the length of the meter by the wavelength of the krypton-86 orange line.

INSTRUMENTS AND CONTROL SYSTEMS

No. 1, 1961

J. Kelly. Electrical measurements and standards laboratories in the USA, Britain, FGR, France and Austria.

Brief description of impressions gained in visits to various national metrological organizations, and in particular their electrical measurements and standards departments. An opinion is expressed on the necessity for exchanging experiences among these organizations.

M. Steinberg. Instrument testing carried out by the Canadian Scientific Research Council.

A description is provided of the functions of the Canadian Scientific Research Council, which are similar to those of the National Bureau of Standards which deals with the testing and certifying of instruments.

J. Riley. Circuits for measuring three- and four- terminal resistors.

Three- and four- terminal resistors and bridge circuits for their measurement are described.

P. Benschoten. Modified "isoohmic" potential dividers characterized by a higher stability and ease of manufacture, and useful for five-decade instruments.

P. Letica. Application of the Schering bridge for measuring capacitance and dielectric loss factors at voltages exceeding one million volts.

G. Stolar. Manufacturing errors. Various types of production errors are described, including errors of electrical circuits, of calibration, of linearity and reproducibility, mainly for direct-reading dc instruments.

J. Yanicke. Test racks for checking grade 0.1 - 3 electrical measuring instruments.

The basic characteristics of test racks for testing electrical measuring instruments are given.

R. Calvert. Inductive ratio-arm bridge.

A bridge is described which is characterized by independent balancing of resistive and reactive components, a wide measuring range and high accuracy.

MESURES & CONTROLE INDUSTRIEL

No. 287, 1961

At the "Mesucora" exhibition (Paris, May 9-17, 1961) measuring, control, regulation and automatic equipment was shown. A list of exhibiting firms is provided with an enumeration and brief description of the instruments exhibited by them.

R. Farden. European production of automatic control instruments.

Review of the present production of automatic control instruments in Europe, published by the French Center of Technical Information.

REVIEW

A. V. ÉRVAIS, ADJUSTMENT AND REPAIR OF MEASURING MACHINES

V. Ya. Éidinov

Mashgiz, Moscow, 1960, 107 pp., 40 drawings.
Edition of 4200 copies, price 34 kopeks.
Translated from Izmeritel'naya Tekhnika, No. 7,
pp. 64, July, 1961

A. V. Érvais' booklet "Adjustment and Repair of Measuring Machines," which consists in essence of the missing chapter of his book "Adjustment and Repair of Opticomechanical Measuring Instruments" (see Izmeritel'naya Tekhnika, No. 6, 1959), is topical and of considerable interest to a wide circle of persons concerned with linear measurements. In this small booklet the author has succeeded in collecting and sorting out extensive material, which includes the latest achievements in designing measuring machines both in the USSR and abroad.

The booklet consists of three chapters. Chapter I reviews the various types of measuring machines manufactured by our own and foreign industries. Some of these machines are mass-produced, others produced individually, but of interest in their design. The description of separate instrument units may be useful for developing other measuring instruments and devices. The relatively large size of this chapter is justified by the exhaustive information on the design, operational characteristics, peculiarities, advantages and disadvantages of the machines under review.

Chapter II provides the limiting errors of the measuring machines and the various components of these errors. These data, as well as the data provided in Chapter I are very useful for selecting and allocating measuring machines.

Chapter III should be considered as the basis of this booklet. The information provided in this chapter is the result of the great practical experience of the author in the adjustment of opticomechanical instruments. Chapter III provides recommendations on efficient methods for dismantling and adjusting machines most commonly used in industry. The methods for testing machines described in this chapter are based on the instructions of the Committee of Standards, Measures and Measuring Instruments, but it also contains some additional suggestions for checking certain machine parameters required for the adjustment of the machines.

The adjustment methods described in the booklet fully meet modern requirements.

Certain adjustments described in the booklet will be difficult for adjuster-mechanics working in unspecialized plants. Certain adjustments (for instance, those of the collimation objectives of tail spindle and measuring heads) can be carried out in a simpler manner. However, the book as a whole fully meets its purpose of providing practical and accurate instructions on the repair and adjustment of measuring machines.

The material is presented in the booklet in an easily understandable manner. The illustrations are adequate in number and of high quality. The appearance of the book is satisfactory, but the drawing on its cover is completely out of place with respect to its content.

The book is undoubtedly useful. It is a pity we lack similar basic handbooks for many other types of measuring instruments.

Soviet Journals Available in Cover-to-Cover Translation

ABBREVIATION	RUSSIAN TITLE	TITLE OF TRANSLATION	PUBLISHER	TRANSLATION BEGAN Vol.	Issue	Year
AÉ	Atomnaya énergiya	Soviet Journal of Atomic Energy	Consultants Bureau	1	1	1956
Akust. zh.	Akusticheskii zhurnal	Soviet Physics – Acoustics	American Institute of Physics	1	1	1955
	Antibiotiki	Antibiotics	Consultants Bureau	4	1	1959
Astr(on). zh(urn).	Astronomicheskii zhurnal	Soviet Astronomy–AJ	American Institute of Physics	34	1	1957
Avto(mat). svarka	Avtomaticheskaya svarka	Automatic Welding	British Welding Research Association (London)		1	1959
	Avtomatika i Telemekhanika	Automation and Remote Control	Instrument Society of America	27	1	1956
	Biofizika	Biophysics	National Institutes of Health*		1	1957
	Biokhimiya	Biochemistry	Consultants Bureau	21	1	1956
Byull. éksp(erim). biol. i med.	Byulleten' éksperimental'noi biologii i meditsiny	Bulletin of Experimental Biology and Medicine	Consultants Bureau	41	1	1959
DAN (SSSR) / Dokl(ady) AN SSSR	Doklady Akademii Nauk SSSR	The translation of this journal is published in sections, as follows:				
	Life Sciences	Doklady Biochemistry Section	American Institute of Biological Sciences	106	1	1956
	Life Sciences	Doklady Biological Sciences Sections (Includes: Anatomy, biophysics, cytology, ecology, embryology, endocrinology, evolutionary morphology, genetics, histology, hydrobiology microbiology, morphology, parasitology, physiology, zoology sections)	American Institute of Biological Sciences	112	1	1957
	Life Sciences	Doklady Botanical Sciences Sections (Includes: Botany, phytopathology, plant anatomy, plant ecology, plant embryology, plant physiology, plant morphology sections)	American Institute of Biological Sciences	112	1	1957
	Chemical Sciences	Proceedings of the Academy of Sciences of the USSR, Section: Chemical Technology	Consultants Bureau	106	1	1956
	Chemical Sciences	Proceedings of the Academy of Sciences of the USSR, Section: Chemistry	Consultants Bureau	106	1	1956
	Chemical Sciences	Proceedings of the Academy of Sciences of the USSR, Section: Physical Chemistry	Consultants Bureau	112	1	1957
	Earth Sciences	Doklady Earth Sciences Sections (Includes: Geochemistry, geology, geophysics, hydrogeology, mineralogy, paleontology, petrography, permafrost sections)	American Geological Institute	124	1	1959
	Earth Sciences	Proceedings of the Academy of Sciences of the USSR, Section: Geochemistry	Consultants Bureau	106-123	1-6	1957-1958
	Earth Sciences	Proceedings of the Academy of Sciences of the USSR, Sections: Geology	Consultants Bureau	106-123	1-6	1957-1958
	Mathematics	Doklady Soviet Mathematics	The American Mathematics Society	131	1	1961
	Physics	Soviet Physics–Doklady (Includes: Aerodynamics, astronomy, crystallography, cybernetics and control theory, electrical engineering, energetics, fluid mechanics, heat engineering, hydraulics, mathematical physics, mechanics, physics, technical physics, theory of elasticity sections)	American Institute of Physics	106	1	1956
	Physics	Proceedings of the Academy of Sciences of the USSR, Applied Physics Sections (does not include mathematical physics or physics sections)	Consultants Bureau	106-117	1	1956-1957
Derevoobrabat. prom-st'.	Derevoobrabatyvayushchaya promyshlennost'	Wood Processing Industry	Timber Development Association (London)		9	1959
	Élecktrosvyaz	Telecommunications	Massachusetts Institute of Technology*		1	1957
Éntom(ol). oboz(renie)	Éntomologicheskoe obozrenie	Entomological Review	American Institute of Biological Sciences	38	1	1959
Farmakol. (i) toksikol(ogiya)	Farmakologiya i toksikologiya	Pharmacology and Toxicology	Consultants Bureau	20	1	1957
FMM	Fizika metallov i metallovedenie	Physics of Metals and Metallography	Acta Metallurgica*	5	1	1957
Fiziol. zhurn. SSSR (im. Sechenova)	Fiziologicheskii zhurnal im. I. M. Sechenova	Sechenov Physiological Journal USSR	National Institutes of Health*		1	1957
Fiziol(ogiya) rast.	Fiziologiya rastenii	Plant Physiology	American Institute of Biological Sciences	4	1	1957
	Geokhimiya	Geochemistry	The Geochemical Society		1	1958
FTT	Fizika tverdogo tela	Soviet Physics–Solid State	American Institute of Physics	1	1	1959
Izmerit. tekh(nika)	Izmeritel'naya tekhnika	Measurement Techniques	Instrument Society of America		1	1959
Izv. AN SSSR, O(td). Kh(im). N(auk)	Izvestiya Akademii Nauk SSSR: Otdelenie khimicheskikh nauk	Bulletin of the Academy of Sciences of the USSR: Division of Chemical Sciences	Consultants Bureau		1	1952

continued

Izv. AN SSSR, O(td). T(ekhn). N(auk): Met(all). i top.	(see Met. i top.)				
Izv. AN SSSR Ser. fiz(ich).	Izvestiya Akademii Nauk SSSR: Seriya fizicheskaya	Bulletin of the Academy of Sciences of the USSR: Physical Series	Columbia Technical Translations	1	1954
Izv. AN SSSR Ser. geofiz.	Izvestiya Akademii Nauk SSSR: Seriya geofizicheskaya	Bulletin (Izvestiya) of the Academy of Sciences USSR: Geophysics Series	American Geophysical Union	1	1954
Izv. AN SSSR Ser. geol.	Izvestiya Akademii Nauk SSSR: Seriya geologicheskaya	Izvestiya of the Academy of Sciences of the USSR: Geologic Series	American Geological Institute	1	1958
Kauch. i rez.	Kauchuk i rezina	Soviet Rubber Technology	Research Association of British Rubber Manufacturers	18	1 1959
	Kinetika i kataliz	Kinetics and Catalysis	Consultants Bureau	1	3 1960
	Koks i khimiya	Coke and Chemistry USSR	Coal Tar Research Association (Leeds, England)		1 1958
Kolloidn. zh(urn).	Kolloidnyi zhurnal	Colloid Journal	Consultants Bureau	14	1 1952
	Kristallografiya	Soviet Physics – Crystallography	American Institute of Physics	2	1 1957
Metalov. i term. obrabot. metal.	Metallovedenie i termicheskaya obrabotka metallov	Metal Science and Heat Treatment of Metals	Acta Metallurgica	6	1 1958
	Metallurg	Metallurgist	Acta Metallurgica		1 1957
Met. i top.	Metallurgiya i topliva	Russian Metallurgy and Fuels	Eagle Technical Publications		1 1960
Mikrobiol.	Mikrobiologiya	Microbiology	American Institute of Biological Sciences	26	1 1957
OS	Optika i spektroskopiya	Optics and Spectroscopy	American Institute of Physics		6 1959
	Pochvovedenie	Soviet Soil Science	American Institute of Biological Sciences		1 1958
	Priborostroenie	Instrument Construction	British Scientific Instrument Research Association		1 1959
Pribory i tekhn. éks(perimenta)	Pribory i tekhnika éksperimenta	Instruments and Experimental Techniques	Instrument Society of America		1 1957
Prikl. matem. i mekh.	Prikladnaya matematika i mekhanika	Applied Mathematics and Mechanics	American Society of Mechanical Engineers		1 1958
PTÉ	(see Pribory i tekhn. éks.)				
	Problemy Severa	Problems of the North	National Research Council of Canada		
Radiotekh.	Radiotekhnika	Radio Engineering	Massachusetts Institute of Technology*	12	1 1957
Radiotekh. i élektronika	Radiotekhnika i élektronika	Radio Engineering and Electronics	Massachusetts Institute of Technology*	2	1 1957
	Stanki i instrument	Machines and Tooling	Production Engineering Research Assoc.		1 1959
	Stal'	Stal (In English)	Iron and Steel Institute		1 1959
Stek. i keram.	Steklo i keramika	Glass and Ceramics	Consultants Bureau	13	1 1956
Svaroch. proiz-vo	Svarochnoe proizvodstvo	Welding Production	British Welding Research Association		4 1959
Teor. veroyat. i prim.	Teoriya veroyatnostei i ee primenenie	Theory of Probability and Its Applications	Society for Industrial and Applied Mathematics		1 1956
Tsvet. Metally	Tsvetnye metally	Nonferrous Metals	Primary Sources		1 1960
UFN	Uspekhi fizicheskikh Nauk	Soviet Physics – Uspekhi (partial translation)	American Institute of Physics	66	1 1958
UKh	Uspekhi khimii	Russian Chemical Reviews	The Chemical Society (London)		1 1960
UMN	Uspekhi matematicheskikh nauk	Russian Mathematical Surveys	London Mathematical Society	15	1 1960
Usp. fiz. nauk	(see UFN)				
Usp. khim(ii)	(see UKh)				
Usp. matem. nauk	(see UMN)				
Usp. sovr. biol.	Uspekhi sovremennoi biologii	Russian Review of Biology	Oliver and Boyd		48 1959
Vest. mashinostroeniya	Vestnik mashinostroeniya	Russian Engineering Journal	Production Engineering Research Assoc.		4 1959
Vop. gem. i per. krovi	Voprosy gematologii i perelivaniya krovi	Problems of Hematology and Blood Transfusion	National Institutes of Health*		1 1957
Vop. onk.	Voprosy onkologii	Problems of Oncology	National Institutes of Health*		1 1957
Vop. virusol.	Voprosy virusologii	Problems of Virology	National Institutes of Health*		1 1957
Zav(odsk). lab(oratoriya)	Zavodskaya laboratoriya	Industrial Laboratory	Instrument Society of America	25	1 1959
ZhAKh Zh. anal(it). khimii	Zhurnal analiticheskoi khimii	Journal of Analytical Chemistry USSR	Consultants Bureau	7	1 1952
ZhETF Zh. éksperim. i teor. fiz.	Zhurnal éksperimental'noi i theoreticheskoi fiziki	Soviet Physics–JETP	American Institute of Physics	28	1 1955
ZhFKh Zh. fiz. khimii	Zhurnal fizicheskoi khimii	Russian Journal of Physical Chemistry	The Chemical Society (London)		7 1959
ZhMÉI Zh(urn). mikrobiol. épidemiol. i immunobiol.	Zhurnal mikrobiologii, épidemiologii i immunobiologii	Journal of Microbiology, Epidemiology and Immunobiology	National Institutes of Health*		1 1957
ZhNKh Zh(urn). neorgan(ich). khim(ii)	Zhurnal neorganicheskoi khimii	The Russian Journal of Inorganic Chemistry	The Chemical Society (London)		1 1959
ZhOKh Zh(urn). obshch(ei) khimii	Zhurnal obshchei khimii	Journal of General Chemistry USSR	Consultants Bureau	19	1 1949
ZhPKh Zh(urn). prikl. khimii	Zhurnal prikladnoi khimii	Journal of Applied Chemistry USSR	Consultants Bureau	23	1 1950
ZhSKh Zh(urn). strukt. khimii	Zhurnal strukturnoi khimii	Journal of Structural Chemistry	Consultants Bureau	1	1 1960
ZhTF Zh(urn). tekhn. fiz.	Zhurnal teknicheskoi fiziki	Soviet Physics–Technical Physics	American Institute of Physics	26	1 1956
Zh(urn). vyssh. nervn. deyat. (im. Pavlova)	Zhurnal vysshei nervnoi deyatel'nosti (im. I. P. Pavlova)	Pavlov Journal of Higher Nervous Activity	National Institutes of Health*		1 1958

*Sponsoring organization. Translation through 1960 issues is a publication of Pergamon Press.

SIGNIFICANCE OF ABBREVIATIONS MOST FREQUENTLY ENCOUNTERED IN SOVIET TECHNICAL PERIODICALS

AN SSSR	*Academy of Sciences, USSR*
FIAN	*Physics Institute, Academy of Sciences USSR*
GITI	*State Scientific and Technical Press*
GITTL	*State Press for Technical and Theoretical Literature*
GOI	*State Optical Institute*
GONTI	*State United Scientific and Technical Press*
Gosénergoizdat	*State Power Press*
Gosfizkhimizdat	*State Physical Chemistry Press*
Goskhimizdat	*State Chemistry Press*
GOST	*All-Union State Standard*
Gostekhizdat	*State Technical Press*
GTTI	*State Technical and Theoretical Press*
IAT	*Institute of Automation and Remote Control*
IF KhI	*Institute of Physical Chemistry Research*
IFP	*Institute of Physical Problems*
IL	*Foreign Literature Press*
IPF	*Institute of Applied Physics*
IPM	*Institute of Applied Mathematics*
IREA	*Institute of Chemical Reagents*
ISN (Izd. Sov. Nauk)	*Soviet Science Press*
IYap	*Institute of Nuclear Studies*
Izd	*Press (publishing house)*
LÉTI	*Leningrad Electrotechnical Institute*
LFTI	*Leningrad Institute of Physics and Technology*
LIM	*Leningrad Institute of Metals*
LITMiO	*Leningrad Institute of Precision Instruments and Optics*
Mashgiz	*State Scientific-Technical Press for Machine Construction Literature*
MGU	*Moscow State University*
Metallurgizdat	*Metallurgy Press*
MOPI	*Moscow Regional Pedagogical Institute*
NIAFIZ	*Scientific Research Association for Physics*
NIFI	*Scientific Research Institute of Physics*
NIIMM	*Scientific Research Institute of Mathematics and Mechanics*
NIKFI	*Scientific Institute of Motion Picture Photography*
NKTM	*People's Commissariat of the Heavy Machinery Industry*
Obrongiz	*State Press of the Defense Industry*
OIYaI	*Joint Institute of Nuclear Studies*
ONTI	*United Scientific and Technical Press*
OTI	*Division of Technical Information*
OTN	*Division of Technical Science*
RIAN	*Radium Institute, Academy of Sciences of the USSR*
SPB	*All-Union Special Planning Office*
Stroiizdat	*Construction Press*
URALFTI	*Ural Institute of Physics and Technology*
TsNIITMASh	*Central Scientific Research Institute of Technology and Machinery*
VNIIM	*All-Union Scientific Research Institute of Metrology*

NOTE: Abbreviations not on this list and not explained in the translation have been transliterated, no further information about their significance being available to us — *Publisher.*

Publication of a "Soviet Instrumentation and Control Translation Series" by the Instrument Society of America has been made possible by a grant in aid from the National Science Foundation, with additional assistance from the National Bureau of Standards for the journal *Measurement Techniques.*

Subscription rates have been set at modest levels to permit widest possible distribution of these translated journals.

The Series now includes four important Soviet instrumentation and control journals. The journals included in the Series, and the subscription rates for the translations, are as follows:

MEASUREMENT TECHNIQUES – *Izmeritel'naya Tekhnika*

Russian original published by the Committee of Standards, Measures, and Measuring Instruments of the Council of Ministers, USSR. The articles in this journal are of interest to all who are engaged in the study and application of fundamental measurements. Both 1958 (bimonthly) and 1959-1961 (monthlies) available.

Per year (12 issues) starting with 1961, No. 1

General:	United States and Canada . .	$25.00
	Elsewhere	28.00
Libraries of nonprofit academic institutions:		
	United States and Canada . .	$12.50
	Elsewhere	15.50

INSTRUMENTS AND EXPERIMENTAL TECHNIQUES
Pribory i Tekhnika Éksperimenta

Russian original published by the Academy of Sciences, USSR. The articles in this journal relate to the function, construction, application, and operation of instruments in various fields of experimentation. 1958-1961 issues available.

Per year (6 issues) starting with 1961, No. 1

General:	United States and Canada . .	$25.00
	Elsewhere	28.00
Libraries of nonprofit academic institutions:		
	United States and Canada . .	$12.50
	Elsewhere	15.50

AUTOMATION AND REMOTE CONTROL – *Avtomatika i Telemekhanika*

Russian original published by the Institute of Automation and Remote Control of the Academy of Sciences, USSR. The articles are concerned with analysis of all phases of automatic control theory and techniques. 1957-1961 1959, and 1960

Per year (12 issues) starting with Vol. 22, No. 1

General:	United States and Canada . .	$35.00
	Elsewhere	38.00
Libraries of nonprofit academic institutions:		
	United States and Canada . .	$17.50
	Elsewhere	20.50

INDUSTRIAL LABORATORY – *Zavodskaya Laboratoriya*

Russian original published by the Ministry of Light Metals, USSR. The articles in this journal relate to instrumentation for analytical chemistry and to physical and mechanical methods of materials research and testing. 1958-1961 issues available.

Per year (12 issues) starting with Vol. 27, No. 1

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